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- Switched-mode power supply transformer design nomograms
- 50W multiple-output switched-mode power supply
- CL8880 series radar traffic sensors

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Switched-mode power supply transformer design nomograms *F. J. Burgum* 354

50W multiple-output switched-mode power supply *M. C. Basell* 379

Product Information: CL8880 series radar traffic sensors 388

Cover picture:
The CL8880B doppler radar module, mounted above a set of portable traffic signals, responds to the approach of a vehicle.

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Switched-mode power supply transformer design nomograms

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These nomograms constitute an aid to ferrite core selection and winding design for push-pull, forward, and flyback converter transformers over a wide range of frequencies and operating conditions.

INTRODUCTION

The nomograms included in this publication provide a ready guide to the design of output transformers for switched-mode power supplies, using the EC series of ferrite cores developed specifically for this application. Transformer design for push-pull, forward, or flyback converter circuits is simplified, and exact results are achieved without the need for specialist knowledge.

In the following pages the use of the nomograms is described in detail, with some theoretical background. Quick-reference instructions are given (in Appendices 1, 2, and 3), and the nomograms have model solutions drawn on them; it is nevertheless suggested that until the user is familiar with the procedure, the full text be consulted each time the nomograms are used. For flyback converter transformers, where the design procedure is necessarily more complex than for the other types, reference to the text is particularly important.

The nomograms are listed in Table 1. Where no statement is made to the contrary, the nomograms are equally applicable to push-pull, forward, and flyback converter transformers. All the symbols used in this publication are listed and explained in Table 2.

ASSUMPTIONS AND SCALING RULES

The nomograms apply to switched-mode power supply transformers using pairs of cores from the EC series of ferrite E-cores under the given conditions. Scaling rules are given for cases where certain of the assumptions are not fulfilled.

Converter operation

The converter operates at a fixed frequency f, and has a voltage-feedback regulation system which adjusts the duty cycle δ to maintain a constant output voltage V_o.

Creepage and clearance distance

A 4mm gap is left on each side of the winding for safety isolation and creepage distance, as shown in Fig.1.

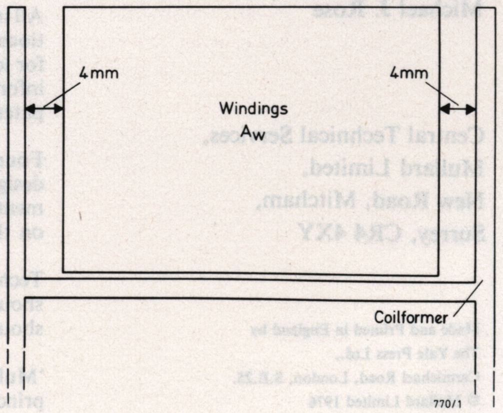


Fig.1 — Safety isolation and creepage distance

TABLE 1
Nomograms Included in this Publication

1a to 1d	Preliminary core selection
2a and 2b	Parameter Δ (flyback converters only)
3	Winding turns per volt
4	Total permitted transformer losses for a range of core centre-pole temperature rises
5	Core loss
6	Winding loss (total transformer loss minus core loss)
7	Throughput power as a function of winding loss
8	Winding inductance (push-pull and forward converters only)
9	Primary peak magnetising current (push-pull and forward converters only)
10	Minimum output power for good regulation (push-pull converters only)
11	Transformer secondary inductance for required regulation range (flyback converters only)
12	Core spacer thickness (flyback converters only)
13	Copper wire outside diameter for minimum loss
14	Copper wire outside diameter for minimum loss (secondaries of split winding only)
15	Copper strip thickness for minimum loss

TABLE 2
Symbols and Units Used in this Publication

Symbol	Quantity	Unit
A_{Cu}	Total copper area in transformer	mm²
A_L	Winding inductance for one turn	nH
A_W	Total available winding area	mm²
f	Converter operating frequency	kHz
F_R	Ratio of a.c. resistance to d.c. resistance of a winding	—
F_W	Winding copper factor (A_{Cu}/A_W)	—
$\hat{I}_{mag}$	Peak primary magnetising current	A
I_o	Converter d.c. output current	A
ℓ_s	Thickness of spacer between core halves for a flyback converter transformer	mm
L_p	Transformer primary inductance	μH
L_s	Transformer secondary inductance	μH
n	Number of turns in a winding	—
n/V	Number of turns per volt in a transformer	V^{-1}
P_C	Transformer core loss	W
P_o	Transformer throughput power	W
P_T	Total transformer losses	W
$P_{tr(min)}$	Minimum transformer throughput power for good regulation in a push-pull converter	W
P_W	Transformer winding loss	W
$R_{L(min)}$	Flyback converter transformer minimum load resistance at maximum output current = $(V_o + V_R)/I_{o(max)}$	Ω
T_a	Ambient temperature	°C
$T_{cp(max)}$	Maximum permissible centre-pole temperature	°C
V_i	Converter d.c. input voltage	V
V_o	Converter d.c. output voltage	V
V_R	Output rectifier voltage drop	V
α	Ratio $\hat{\Phi}_{max}/\hat{\Phi}$, usually equal to $V_{i(max)}/V_{i(min)}$	—
δ	Duty cycle (maximum value 0.5 for push-pull converters)	—
Δ	Ratio $I_{o(min)}/I_{o(max)}$ at $\delta_{min} = \delta_{max} = 0.5$ for a flyback converter	—
$\Delta T_{(cp\text{-}a)}$	Transformer core centre-pole temperature rise above ambient	degC
μ_a	Relative amplitude permeability for a pair of EC cores with no air gap	—
$\hat{\Phi}$	Core peak flux under normal steady-state conditions	Wb
$\hat{\Phi}_{max}$	Maximum permissible peak flux in core. For the EC series of cores with a maximum centre-pole temperature of 100°C, this is the flux corresponding to a flux density of 0.32T in a centre-pole of minimum specified cross-sectional area	Wb
$\hat{\Phi}_{trans}$	Core peak flux under transient worst-case conditions	Wb

If the 4mm clearances are omitted (that is, if the winding occupies the entire width of the coil former), then the transformer throughput power P_o obtained from Nomogram 7 must be modified. The modified throughput power P' is given by:

$$P_o{}' = K_i P_o,$$

where the factor K_i is as given in Table 3 (Page 371).

Core temperature
The maximum core centre-pole temperature $T_{cp(max)}$ is 100°C.

Ancillary windings
All windings contribute to throughput power.

If the coil former carries any ancillary windings not contributing to throughput power and having negligible dissipation, such as an energy-recovery winding, then the throughput power obtained from Nomogram 7 must again be modified, as follows:

$$P_o{}' = P_o \times \sqrt{\left(\frac{\text{area occupied by useful power windings}}{\text{total area occupied by windings}}\right)}.$$

For example, if a forward converter transformer contains a primary, a secondary, and an energy-recovery winding, each taking up approximately equal space on the coil former, then because the energy-recovery winding does not contribute to throughput power we have:

$$\frac{\text{area occupied by useful power windings}}{\text{total area occupied by windings}} = \frac{2}{3}.$$

Therefore,

$$P_o{}' = P_o \times \sqrt{(2/3)}.$$

The above example (that is, the allowance for an energy-recovery winding in a forward converter transformer) is given as an illustration of the technique. In this particular case, however, the calculation will not be necessary, because for forward and flyback converters both '1+1' and '2+1' transformers are catered for in the nomograms. In Nomogram 7 the use of the appropriate point on the converter type scale, or (for a flyback converter) the appropriate Δ scale, will give the correct value of P_o.

If further ancillary windings are included in either forward, flyback, or push-pull converter transformers, then they may be taken into account by using the equation given above.

Maximum duty cycle
The maximum duty cycle of the converter is $\delta_{max} = 0.5$.

For forward and push-pull converters with δ_{max} other than 0.5, the modified throughout power is:

$$P_o{}' = \frac{P_o}{\sqrt{(2\,\delta_{max})}}.$$

The number of turns per volt (n/V) obtained from Nomogram 3 should also be modified. The new value is:

$$(n/V)' = (n/V) \times 2\,\delta_{max},$$

for all primary windings, and for the secondaries of push-pull and forward converter transformers; and:

$$(n/V)' = (n/V) \times \frac{\delta_{max}}{1-\delta_{max}},$$

for flyback converter transformer secondary windings.

USE OF THE NOMOGRAMS

Table 4 lists the information required to start a design. The second item in the list (transformer winding configuration) perhaps requires some further explanation. In this context, 'two primary windings' means a double winding only half of which is conducting at any time: a centre-tapped winding is included in this definition. Similarly, 'two secondaries' means a secondary winding only half of which is conducting at any time. Circuit symbols illustrating the terms '1+1', '2+1', and '2+2' are given in Fig.2.

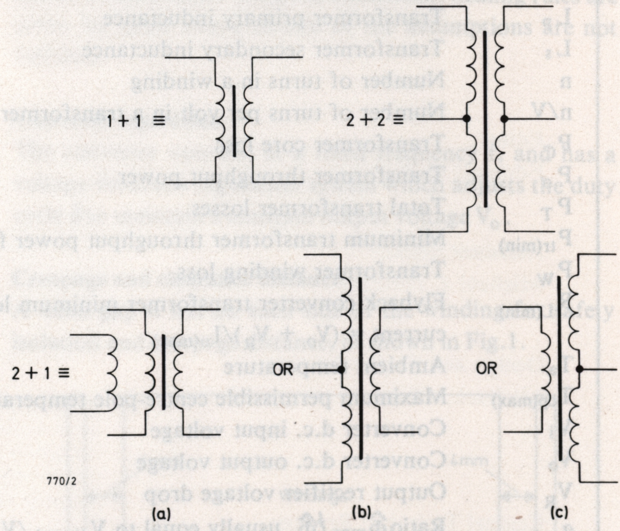


Fig.2 — Explanation of the terms '1+1', '2+1', and '2+2'. The '1+1' configuration is applicable to a forward, flyback, or push-pull converter transformer. The '2+2' configuration, and the '2+1' (b) and (c) configurations, are applicable to a push-pull converter transformer. The '2+1' (a) configuration is applicable to a forward converter transformer with an energy-recovery winding, and to a flyback converter transformer with a voltage-limiting winding. In both cases it is assumed that the ancillary winding occupies one-third of the available winding space and has negligible ddisspiation.

TABLE 4
Information Required from the User

Quantity	Symbol	Unit
Type of converter circuit	—	—
Transformer winding configuration (number of windings each taking approximately equal space on the coil former — see Fig.2):		
one primary and one secondary	1+1	—
one primary and two secondaries, or two primaries and one secondary	2+1	—
two primaries and two secondaries	2+2	—
Maximum d.c. input voltage	$V_{i(max)}$	V
Minimum d.c. input voltage	$V_{i(min)}$	V
D.C. output current	I_o	A
D.C. output voltage	V_o	V
Output rectifier voltage drop	V_R	V
Operating frequency	f	kHz
Maximum permissible core centre-pole temperature rise above ambient*	$\Delta T_{(cp\text{-}a)}$	degC
For flyback converters only:		
maximum duty cycle	δ_{max}	—
minimum output current for which good load regulation (constant duty cycle) is to be maintained	$I_{o(min)}$	A

*Most of the nomograms assume $T_{cp(max)} = 100°C$. Thus the user is to supply T_a and then derive $\Delta T_{(cp\text{-}a)}$.

Nomogram 1 — Preliminary core selection
Nomograms 1a to 1d are for use in making an initial estimate of the type of core necessary for the transformer. The following assumptions are made.

1) The core centre-pole temperature rise above ambient, $\Delta T_{(cp\text{-}a)}$, is 40 deg C.
If a different centre-pole temperature rise is required, then the modified throughput power is:

$$P_o{}' = P_o \times \sqrt{\left(\frac{\Delta T_{(cp-a)}}{40}\right)} = P_o \times \sqrt{\left(\frac{100 - T_a}{40}\right)}.$$

2) The ratio $F_W / F_R = 0.4$ (see Table 2).
3) The value of α is determined (see below).
4) A clearance of 4mm is allowed at each end of the winding.

Before using the nomograms, it is necessary to choose the type of converter circuit to be employed. Detailed information on push-pull, forward, and flyback converter circuits is given in Ref. 1.

Nomograms 1a and 1b apply to push-pull converter transformers. Nomogram 1a has:

$$\alpha = \frac{\hat{\Phi}_{max}}{\hat{\Phi}} \geqslant 1{\cdot}72,$$

and makes no allowance for any d.c. flux component in the core due to unbalanced push-pull operation; if an imbalance exists, then the core flux under transient worst-case conditions could exceed $\hat{\Phi}_{max}$.

The value of α determines the safety margin available in the core for handling transient effects without saturation.

For a forward or push-pull converter in normal operation, the feedback loop adjusts the duty cycle to keep the output voltage constant. Thus, $V_o \propto \delta V_i$ (independent of load current within the control range), and therefore $\delta \propto 1/V_i$. But the peak core flux is proportional to δV_i, and, since this quantity is kept constant, the peak flux is independent of input voltage. Thus:

$$\delta V_i = V_{i(max)}\,\delta_{min} = V_{i(min)}\,\delta_{max},$$
$$\propto \hat{\Phi},$$

and $\hat{\Phi}$ is therefore constant.

However, under certain transient conditions such as a sudden increase in load at the instant when $V_i = V_{i(max)}$, it is possible to have $V_i = V_{i(max)}$ and $\delta = \delta_{max}$ simultaneously. In this case the transient peak flux is:

$$\hat{\Phi}_{trans} \propto V_{i(max)}\,\delta_{max};$$

that is:

$$\frac{\hat{\Phi}_{trans}}{\hat{\Phi}} = \frac{V_{i(max)}}{V_{i(min)}}.$$

It is usual to choose $\hat{\Phi}_{trans}$ near the limit of core saturation, with the usual safety margins; that is, $\hat{\Phi}_{trans} = \hat{\Phi}_{max}$, in which case:

$$\alpha = \frac{\hat{\Phi}_{max}}{\hat{\Phi}} = \frac{\hat{\Phi}_{trans}}{\hat{\Phi}} = \frac{V_{i(max)}}{V_{i(min)}}.$$

In mains-fed converters, after allowing for mains drop-out and ripple:
for 220V mains, $V_{i(max)} = 340V$ d.c., $V_{i(min)} = 198V$ d.c.;
for 240V mains, $V_{i(max)} = 371V$ d.c., $V_{i(min)} = 216V$ d.c.;
and so in both cases if $\hat{\Phi}_{trans} = \hat{\Phi}_{max}$ then $\alpha = 1.72$ is used.

If the core losses are too large, or in cases where a push-pull transformer has some imbalance, then to give an improved safety margin $\hat{\Phi}_{trans} < \hat{\Phi}_{max}$ is used. In Nomogram 1b the value chosen is $\alpha = 3.44$, giving $\hat{\Phi}_{trans} = \frac{1}{2}\hat{\Phi}_{max}$ as an example of a design where some imbalance may exist. Nomogram 1b allows for a d.c. component of the flux of up to $\frac{1}{2}\hat{\Phi}_{max}$.

Nomograms 1c and 1d are for forward and flyback converter transformers respectively. Nomogram 1d includes the parameter Δ, a measure of the range of converter d.c. output currents for which the duty cycle δ is independent of load current. A value of $\Delta = 1$ corresponds to a 'ringing-choke' type of flyback converter in which the core flux falls to zero during each cycle. The value $\Delta = 0.25$ corresponds to a flyback converter in which the flux does not fall to zero under full load conditions; the flux will fall to zero if the d.c. load current is reduced to 25% of its maximum value at minimum supply voltage, or to 40% of its maximum value at maximum supply voltage. In the range from 100% to 40% of maximum output current, regulation is good because the duty cycle remains approximately constant with changes of load current. The meaning of Δ is explained more fully in the text for Nomogram 2.

Having decided which of Nomograms 1a to 1d applies, choose the appropriate winding configuration: either '1 + 1', '2 + 1', or '2 + 2' (Fig.2).

Select the appropriate nomogram. Find the intersection of the chosen operating frequency and the required output power. The required core size is that corresponding to the intended winding configuration lying immediately *above* the intersection.

Proceed to Nomogram 2 (for a flyback converter) or 3 (for a forward or push-pull converter).

Nomogram 2—Parameter Δ as a function of $I_{o(min)}/I_{o(max)}$

(flyback converters only)

An important parameter of a flyback converter transformer is the range of output currents over which it gives good regulation in circuit. There is a maximum value of output current, which depends on the transformer peak primary current (limited by the transformer saturation flux and its permissible temperature rise); there is also a minimum, or boundary, output current below which the core flux reaches zero at some point during the cycle. If the output current falls below the minimum, then regulation becomes poor or the duty cycle must vary with output current. Between these limits the converter duty cycle is almost independent of the load current, and regulation is good.

For a given value of peak primary current, the d.c. output current is a function of the duty cycle, somewhat as shown by the line marked 'maximum' in Fig.3. The minimum (boundary) output current is a similar function, shown by the 'minimum' curve in Fig.3. The duty cycle can vary over the range:

$$\delta_{min} \leqslant \delta \leqslant \delta_{max},$$

and so the actual value of maximum and minimum converter output current can also vary. The largest and smallest values of output current that the designer can specify are shown in Fig.3 as $I_{o(max)}$, the smallest value of maximum current, at $\delta = \delta_{max}$; and $I_{o(min)}$, the largest value of minimum current, at $\delta = \delta_{min}$.

Because of the large range of values of δ_{min}, δ_{max}, $I_{o(min)}$, and $I_{o(max)}$ which may be specified, it is impractical to include all four parameters in each nomogram for flyback converters. Instead, a new parameter Δ is used, defined as:

$$\Delta = \frac{\text{minimum value of } I_o \text{ at } \delta = 0{\cdot}5}{\text{maximum value of } I_o \text{ at } \delta = 0{\cdot}5}.$$

It is then necessary to use an extra nomogram which relates Δ to $I_{o(min)}$, $I_{o(max)}$, δ_{min}, and δ_{max}: this is the purpose of Nomogram 2. In this nomogram δ_{min} is calculated from δ_{max} and the ratio $V_{i(max)}/V_{i(min)}$, and then Δ is calculated from δ_{min} and $I_{o(min)}/I_{o(max)}$.

Nomogram 2 is divided into two parts. In (a), the minimum duty cycle δ_{min} is obtained from δ_{max} and $V_{i(max)}/V_{i(min)}$; in (b), this value of δ_{min} is used with $I_{o(min)}/I_{o(max)}$ to derive Δ.

Proceed to Nomogram 3.

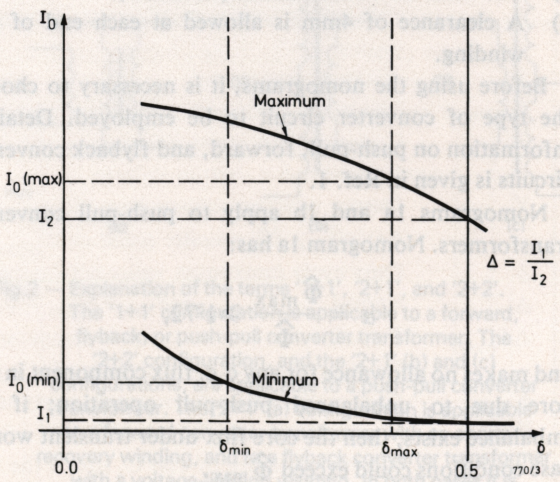


Fig.3 — Flyback converter output current as a function of the duty cycle.

Nomogram 3 — Winding turns per volt

It is now necessary to calculate the number of turns in the primary and secondary windings. The information available comprises the operating frequency, the type of converter circuit, the core size, and the values of α and (for flyback converters) Δ.

On Nomogram 3, draw construction lines joining: core type and value of α; intersection with axis and frequency f. Then join the new intersection with the core type axis and either the converter type or, for a flyback converter, the value of Δ. Read the number of turns per volt given by the intersection of the latter line with the appropriate scale.

The number of turns per volt can then be multiplied by the winding voltages to obtain the number of turns on each winding, as follows. (For secondary windings the rectifier voltage drop V_R must be taken into account.)
Number of primary turns $n_p = V_{i(min)} \times (n/V)$
Number of secondary turns
(flyback or push-pull) $n_s = (V_o + V_R) \times (n/V)$
Number of secondary turns
(forward) $n_s = 2(V_o + V_R) \times (n/V)$
Note that if the maximum duty cycle δ_{max} is not close to 0.5, then it will be necessary to apply the scaling rules for (n/V) quoted on page 356.

The numbers of turns, n_p and n_s, must be rounded to integral values. First find the exact turns ratio $r = n_p/n_s$. Round n_s to the nearest integer *above,* N_s. Calculate the corresponding number of primary turns N_p by rounding rN_s to the nearest integer *below:* this ensures that (for a step-down transformer) the transformer secondary voltage will be sufficiently high.

Check that $N_p \geq n_p$, to ensure that the peak flux in the core has not been increased by the rounding process. If it is found that $N_p < n_p$, then increase N_s by one turn and repeat the calculation for N_p. It may be necessary to repeat this process if either N_p or N_s must be an even number.

The true value of α must now be found. First calculate the true value of (n/V):

$$(n/V) = \frac{N_p}{V_{i(min)}};$$

and then use Nomogram 3 in reverse to find the true value of α. This is the value which must be used in all further steps of the design procedure.

Proceed to Nomogram 4.

Nomogram 4—Total permissible transformer loss

Starting with the value of core centre-pole temperature rise required, $\Delta T_{(cp-a)}$, and the core size (selected from Nomogram 1), the total transformer dissipation P_T is found.

The maximum permissible core centre-pole temperature $T_{cp(max)}$ is 100°C for the value of peak flux $\hat{\Phi}_{max}$ used in these nomograms. Consequently $\Delta T_{(cp-a)}$ is related to the ambient temperature T_a by:

$$\Delta T_{(cp-a)} = 100 - T_a.$$

Proceed to Nomogram 5.

Nomogram 5—Core loss

Draw construction lines joining: frequency and core type; intersection with axis and the chosen value of α. The latter passes through the core loss P_C.

Note that two α scales are provided: one for push-pull converters, and the other for forward and flyback converters.

For a flyback converter the value of P_C must be modified using the parameter Δ. A straight line cuts the values of Δ, P_C, and the new core loss on the P_C (flyback) scale.

Proceed to Nomogram 6.

Nomogram 6—Winding loss and efficiency

The core loss P_C from Nomogram 5 is subtracted from the total transformer loss P_T from Nomogram 4, to give the available transformer winding loss:

$$P_W = P_T - P_C.$$

For push-pull transformers, the design has maximum efficiency if the line joining the values of P_T, P_W, and P_C passes closely above the arrow (which indicates $P_C = 0.8P_W$).

If the line passes below the arrow, the core loss is too high and should be reduced by selecting either a higher value of α or a lower frequency of operation. To do this, draw a line on Nomogram 6 through the value of P_T (from Nomogram 4) and the arrow-head, and read off the values of P_W and P_C. Use Nomogram 5 in reverse, starting from the value of P_C just found to obtain either a new frequency or a new value of α, and then return to Nomogram 3.

Proceed to Nomogram 7.

Nomogram 7 — Transformer throughput power

Draw construction lines joining: converter type and the value of F_W/F_R; core type and frequency f. Then join the intersection with axis 1 and the value of α; the intersection with axis 2 and the intersection with axis 3. The intersection with axis 4 then gives the proportionality between winding loss P_W and throughput power P_o.

For a flyback converter transformer, the value of P_o must be modified. Draw a further line from P_o through the value of Δ appropriate to the winding configuration, to the P_o (flyback) scale.

The parameter F_W is the copper filling factor for the transformer, and F_R is the a.c.-to-d.c. resistance ratio of the windings. For many designs operating in the region of 25kHz, the value of F_W/F_R is in the range from 0.3 to 0.5. Thus a value of $F_W/F_R = 0.4$ may be used to find the *available* transformer throughput power; or the *required* throughput power may be used to find the value of F_W/F_R which must be achieved in practice.

Good estimates of F_W and F_R can be obtained after using Nomogram 13, 14, or 15. To find F_W, first find the total copper area in the transformer:

$$A_{Cu} = \sum_i \frac{\pi n_i d_i^2}{4},$$

where n_i and d_i are the number of turns and copper diameter of the i^{th} winding. (Perform the summation only for windings carrying useful power: do not include energy-recovery windings.) The value of F_W is then given by dividing A_{Cu} by the maximum available winding area of the coil former, calculated from published data and allowing 4mm at either side of the winding as in Fig.1.

The value of F_R is read directly from a scale in each of Nomograms 13 and 14; in Nomogram 15 it has a nearly constant value of 1.33.

For a transformer with no 4mm clearances at each end of the winding, or where the maximum duty cycle is not 0.5, use the scaling rules given on page 356.

The use of Nomogram 7 will confirm either positively or negatively the provisional choice of core size from Nomogram 1. If Nomogram 7 shows that the necessary value of F_W/F_R is impracticably high (greater than, say, 0.6), then it is probably necessary to recommence the design using a larger core. Values of $F_W/F_R < 0.4$ are usually easy to obtain at frequencies of 25kHz or less.

For a push-pull or forward converter transformer, proceed to Nomogram 8. For a flyback converter transformer, proceed to Nomogram 11.

Nomogram 8 — Transformer winding inductance
(push-pull and forward converter transformers only)
If the value of the primary magnetising current is of importance, then Nomogram 8 may be used to find the primary inductance from the number of primary turns and the core size.

Proceed to Nomogram 9.

Nomogram 9 — Primary peak magnetising current
(push-pull and forward converter transformers only)
Draw a construction line joining the frequency f and the minimum input voltage $V_{i(min)}$. A further line joins the intersection with the axis, the primary inductance L_p, and the peak magnetising current $\hat{I}_{mag}$.

For a push-pull converter transformer, proceed to Nomogram 10. For a forward converter transformer, proceed to Nomogram 13, 14, or 15.

Nomogram 10 — Minimum output power for good regulation
(push-pull converter transformers only)
The minimum output power which must be taken from the transformer to achieve good regulation (with an output filter choke having infinite inductance) is found from this nomogram.

Draw a construction line joining frequency f and core type; a further line joins the intersection with the axis, the value of α, and the minimum throughput power $P_{tr(min)}$.

To $P_{tr(min)}$ must be added a component due to the finite size of the output choke:

$$P_{tot(min)} = P_{tr(min)} + P_{ch(min)}.$$

Proceed to Nomogram 13, 14, or 15.

Nomogram 11 — Transformer secondary inductance for good regulation
(flyback converter transformers only)

The value required for the secondary inductance of a flyback converter transformer is found from this nomogram. The starting parameter is $R_{L(min)}$, the minimum load resistance seen by the transformer secondary, defined by:

$$R_{L(min)} = \frac{V_o + V_R}{I_{o(max)}}.$$

This parameter, the operating frequency, and the required value of Δ (from Nomogram 2) are used to find the secondary inductance L_s. The value of L_s obtained ensures that the ratio $I_{o(min)}/I_{o(max)}$ initially required can be achieved at constant duty cycle, and therefore that converter regulation will be good.

Proceed to Nomogram 12.

Nomogram 12 — Core spacer thickness or gap size
(flyback converter transformers only)
The number of secondary turns has been determined from Nomogram 3, and the secondary inductance from Nomogram 11. In Nomogram 12, these two quantities are used to find the thickness of the spacer required in the core for a flyback converter transformer.

Draw a construction line joining the number of turns, the inductance, and the gapped core axis. If the intersection with the latter axis is close to the marked point for the required core type, then this core with a standard gap may be used.

Otherwise, draw a further line from the intersection with the gapped core axis, cutting the required core type on the spaced core axis, and indicating on the spacer thickness scale the required spacer thickness.

Proceed to Nomogram 13, 14, or 15.

Nomograms 13, 14, and 15 — Copper wire diameter (or strip thickness) for minimum loss
These nomograms give the outside diameter (including insulation) of wire, or the thickness of copper strip, with which a winding should be constructed to give the minimum winding loss and hence the minimum transformer temperature rise. Nomograms 13 and 14 are for wire of circular cross-section, and Nomogram 15 is for flat copper strip of width equal to that of the total winding.

The choice of nomogram, and the manner in which it is used, depends on the transformer winding configuration. Three possible winding configurations are shown in Fig.4. A 'simple' winding is one where the whole of one winding is wound on the coil former, followed by the whole of a second winding, as shown in Figs.4a and 4b. The distribution of magnetomotive force (m.m.f.) in such a winding is shown in Fig.5a: the m.m.f. reaches a peak at the junction of the two windings.

A 'split' or 'sandwich' winding has one winding between the halves of another winding (Fig.4c). The distribution of m.m.f. in such a winding (Fig.5b) has two equal peaks of opposite polarity, and (other things being equal) the peak m.m.f. in the split winding is only half that in the simple winding. Because of the induction of eddy currents in the copper windings by the m.m.f. (proximity effect), the split winding with a lower peak m.m.f. has a significantly lower a.c. loss than the simple winding (for further discussion of this concept see Chapter 11 of Ref.2).

A penalty must, however, be paid for the smaller loss given by a split winding. The winding volume and the cost of a split-winding transformer are both increased: a split winding requires twice as many screens (usually four) and more insulation, compared with a simple winding, and the increased number of wire lead-outs costs more to connect. The choice between a split winding (higher cost, lower loss) and a simple winding (lower cost, higher loss) is for the individual designer.

In the calculation of transformer a.c. winding losses, the part of the winding which is important is any portion between surfaces of zero and maximum m.m.f. In a simple winding, a complete primary and a complete secondary each form one winding portion; in a split winding the winding portion is a half-primary or a half-secondary. Note that a half-secondary (but not a half-primary) may contain an odd half layer if the surface of zero m.m.f. cuts a layer of the secondary.

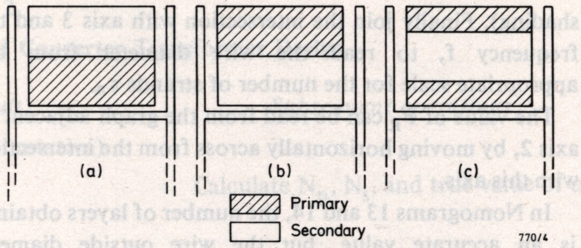


Fig.4 — Winding configurations:
(a) and (b) simple windings
(c) split winding

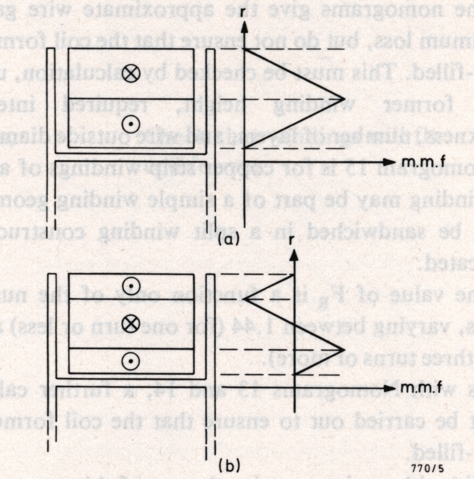


Fig.5 — Distribution of m.m.f. in simple and split windings:
(a) simple winding
(b) split winding

Nomogram 13 is for use with a single winding portion; that is, with a complete primary or secondary of a simple winding, or with a half-primary of a split winding. Nomogram 14 is for use with a double winding portion: the complete secondary of a split winding. Table 5, page 376, can be used to select the appropriate nomogram and the number of turns and layers to use in the nomogram. In Table 5:

n represents the total number of turns in the winding;
m represents the total number of layers in the winding;
n′ is the number of turns to use in the nomogram;
m′ is the number of layers given by the nomogram.

Note that in the case of a split primary the values of n′ and m′ in the nomogram are half those for the whole winding; that is, they are correct for each of the two half-windings.

Nomograms 13 and 14 are used in an identical manner. The labels '$s_n = 1$' and '$s_n = 2$' refer to single and bifilar wire respectively. Use $s_n = 2$ if the winding is bifilar, and if only one of the pair of wires is conducting at any time. Bifilar windings where the pair of wires are connected in parallel are treated as having $s_n = 1$ but twice the true number of turns.

Draw construction lines joining: frequency f (for the appropriate number of strands) and the core type; intersection with axis 1 and number of turns; intersection with axis 2 and the appropriate number of layers (follow shading). Finally join the intersection with axis 3 and the frequency f, to read the wire diameter from the appropriate scale for the number of strands s_n.

The value of F_R can be read from the graph adjacent to axis 2, by moving horizontally across from the intersection with this axis.

In Nomograms 13 and 14, the number of layers obtained is an accurate value, but the wire outside diameter obtained should be treated as a guide only. The predicted value of wire outside diameter will typically be within 10% of the true size required.

The nomograms give the approximate wire gauge for minimum loss, but do not ensure that the coil former is not over-filled. This must be checked by calculation, using the coil former winding height, required interleaving thickness, number of layers, and wire outside diameter.

Nomogram 15 is for copper strip windings of any type. A winding may be part of a simple winding geometry, or may be sandwiched in a split winding construction, as indicated.

The value of F_R is a function only of the number of turns, varying between 1.44 (for one turn or less) and 1.33 (for three turns or more).

As with Nomograms 13 and 14, a further calculation must be carried out to ensure that the coil former is not over-filled.

With either wire or strip, the use of thinner conductors will result in better (lower) values of F_R than predicted by these nomograms; thus if the coil former is over-filled, then a wire diameter should be chosen which will just fill the coil former. However, should the coil former be under-filled, then a thicker conductor should *not* be used as this would result in increased losses.

This completes the transformer design.

ACKNOWLEDGEMENTS

Thanks are due to Mr. L.E. Jansson of Mullard Central Application Laboratory whose work formed the design basis of many of the nomograms, and who contributed many helpful comments during the preparation of this publication; to Mr. M.H. Dryden of Mullard Central Technical Services whose notes formed the basis of the text; and particularly to Mr. A.A.M. Rouwenhorst of Philips' ELCOMA Division, Eindhoven, The Netherlands, who produced the original drawings from which the nomograms have been printed.

REFERENCES

1. JANSSON, L. E. 'A survey of converter circuits for switched-mode power supplies', **Mullard Technical Communications,** Vol. 12, No. 119, July, 1973, pp 271 to 278. Available as reprint TP1442/1.
2. SNELLING, E. C. 'Soft ferrites, properties and applications', Iliffe, London, 1969.

APPENDIX 1

Quick-reference Instructions for a Push-pull Converter Transformer Design

Nomogram	*Information required*	*Result*	*Subsequent operations*
1a or 1b*	f, P_o, winding configuration	Core type (tentative)	–
3	f, estimated a, core type	n/V	Calculate N_p, N_s, and true value of a
4	$\Delta T_{(cp-a)}$, core type	P_T	–
5	f, a, core type	P_C	–
6	P_C, P_T	P_W	Check efficiency. Choose new f or a if necessary (see page 359)
7	P_W, a, f, core type, F_W/F_R	P_o, confirmation of core type	–
8	N_p, core type	L_p	–
9	$V_{i(min)}$, f, L_p	$\hat{I}_{mag}$	–
10	f, a, core type	$P_{tr(min)}$	Add $P_{ch(min)}$
13, 14, 15	N, winding configuration, f, core type	Number of layers, F_R, D_o	Check that windings will fit coil former

*Selected according to the estimated value of a – see page 357.

APPENDIX 2

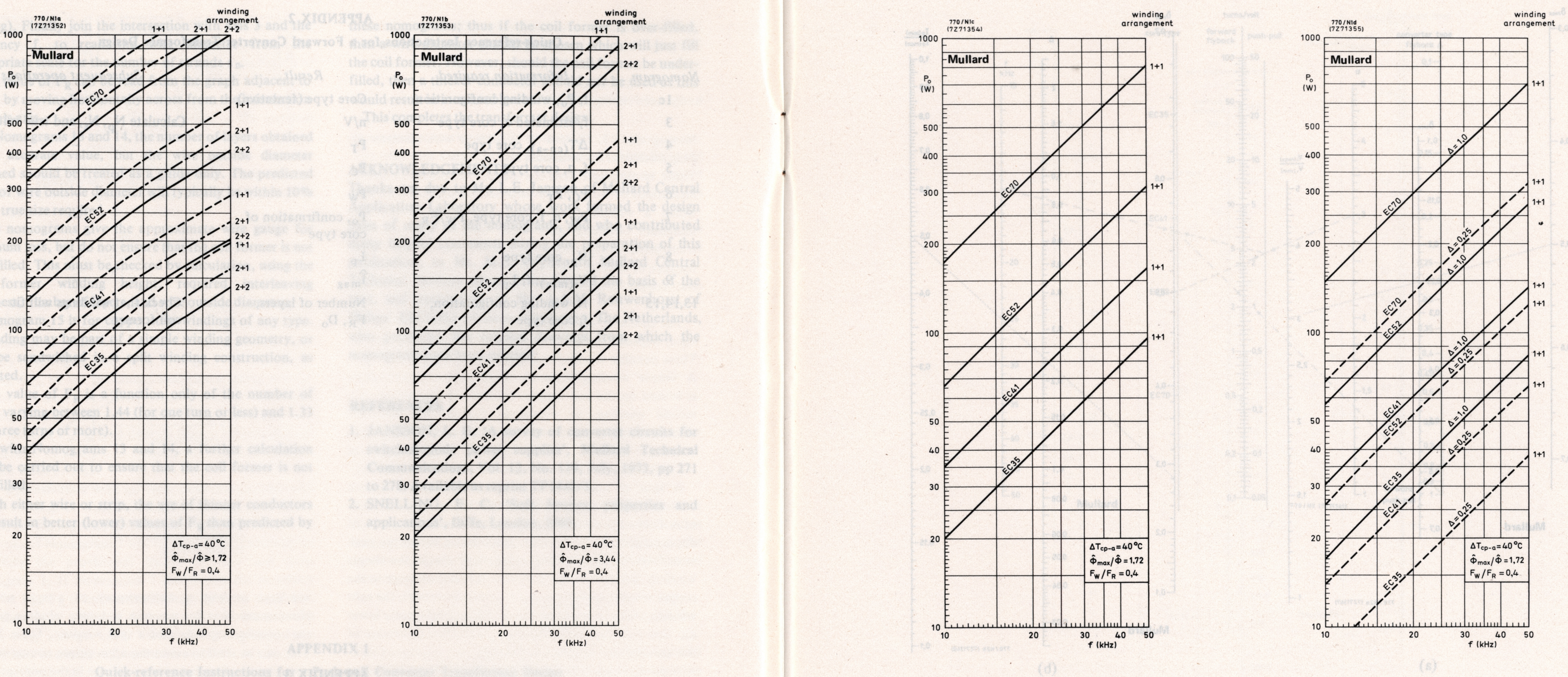
Quick-reference Instructions for a Forward Converter Transformer Design

Nomogram	*Information required*	*Result*	*Subsequent operations*
1c	f, P_o, winding configuration	Core type (tentative)	–
3	f, estimated a, core type	n/V	Calculate N_p, N_s, and true value of a
4	$\Delta T_{(cp-a)}$, core type	P_T	–
5	f, a, core type	P_C	–
6	P_C, P_T	P_W	–
7	P_W, a, f, core type, F_W/F_R	P_o, confirmation of core type	–
8	N_p, core type	L_p	–
9	$V_{i(min)}$, f, L_p	$\hat{I}_{mag}$	–
13,14,15	N, winding configuration, f, core type	Number of layers, F_R, D_o	Check that windings will fit coil former

APPENDIX 3

Quick-reference Instructions for a Flyback Converter Transformer Design

Nomogram	*Information required*	*Result*	*Subsequent operations*
1d	f, P_o, winding configuration, Δ	Core type (tentative)	–
2	$V_{i(max)}/V_{i(min)}$, δ_{max} $I_{o(min)}/I_{o(max)}$	Δ	–
3	f, estimated a, core type, Δ	n/V	Calculate N_p, N_s, and true value of a
4	$\Delta T_{(cp-a)}$, core type	P_T	–
5	f, a, core type, Δ	P_C	–
6	P_C, P_T	P_W	–
7	P_W, a, f, core type, F_W/F_R, Δ	P_o, confirmation of core type	–
11	$R_{L(min)}$, f, Δ	L_s	–
12	N_s, L_s	A_L, ℓ_s	–
13, 14, 15	N, winding configuration, f, core type	Number of layers, F_R, D_o	Check that windings will fit coil former



(a) (b) (c) (d)

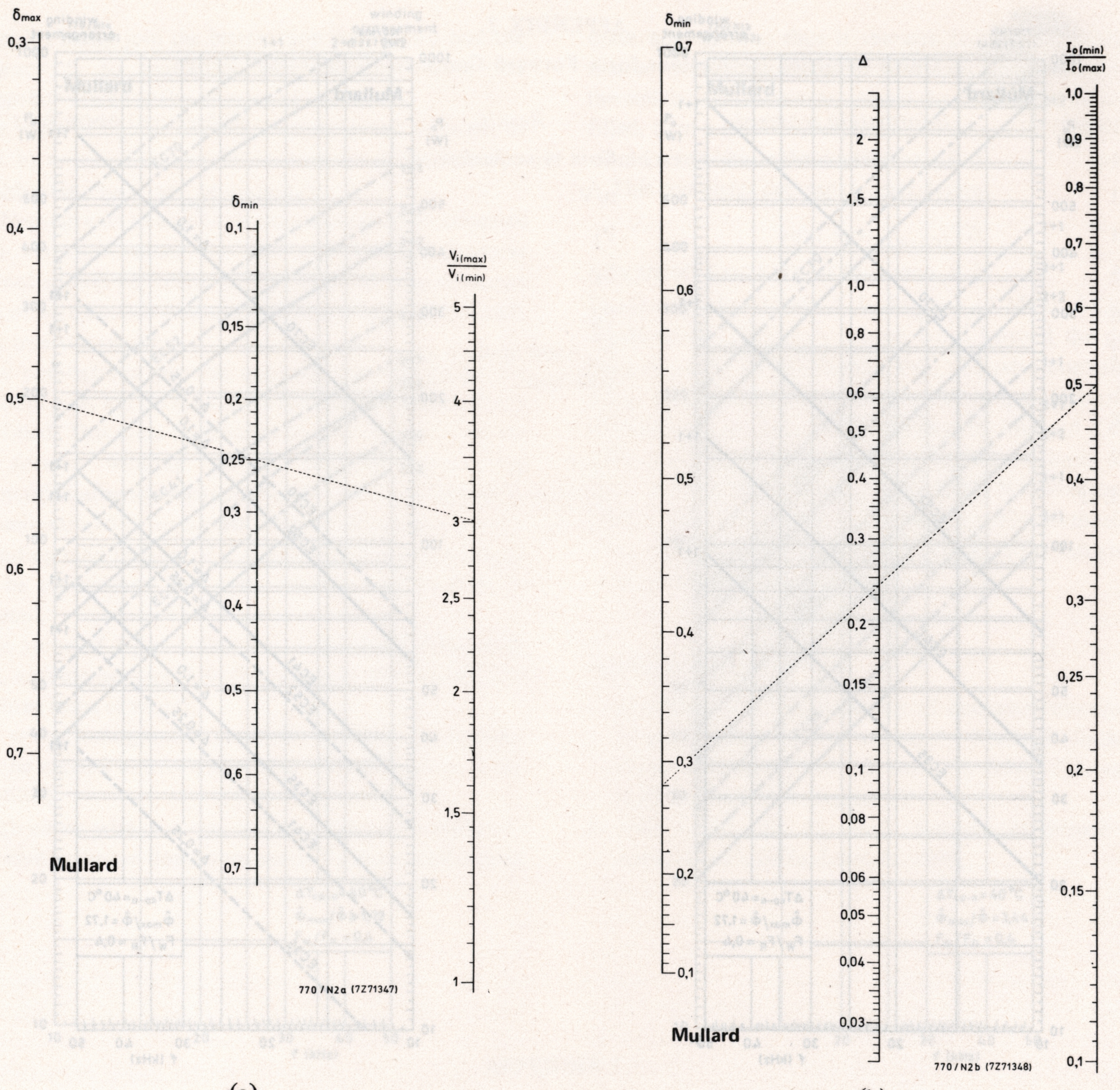
Nomogram 1 — Preliminary core selection

(a) and (b): push-pull converter transformer

(c): forward converter transformer

(d): flyback converter transformer

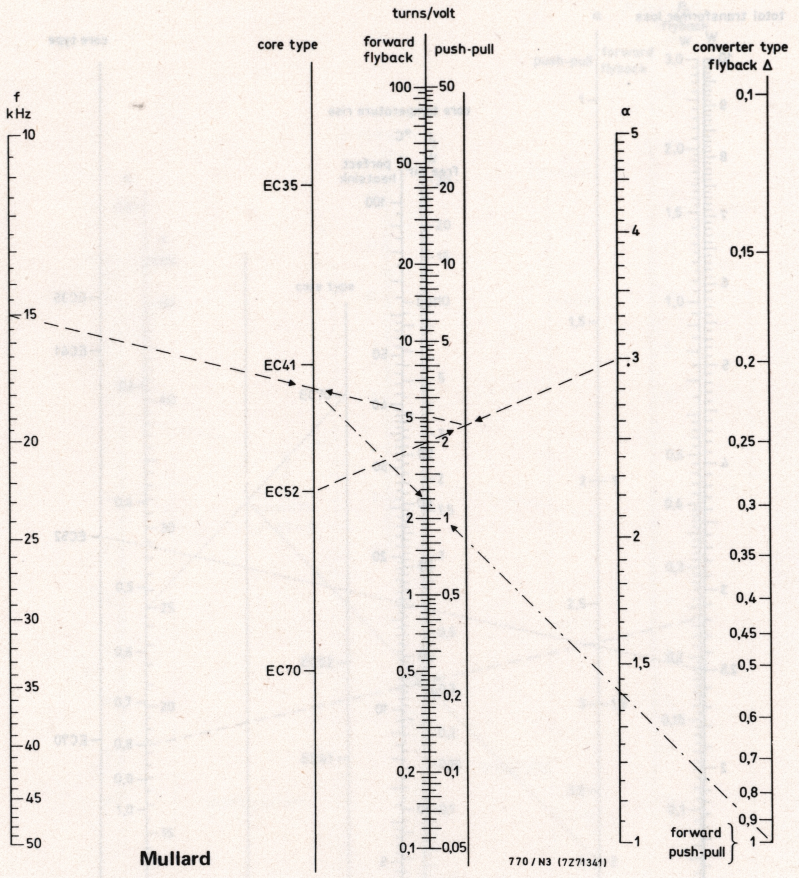
Select the appropriate nomogram. Find the intersection of the chosen operating frequency and the required output power. The required core size is that corresponding to the intended winding configuration lying immediately *above* the intersection.


δmax
δmin
Vi(max)/Vi(min)
Mullard
770 / N2a (7Z71347)
δmin
Δ
Io(min)/Io(max)
Mullard
770 / N2b (7Z71348)

(a) (b)

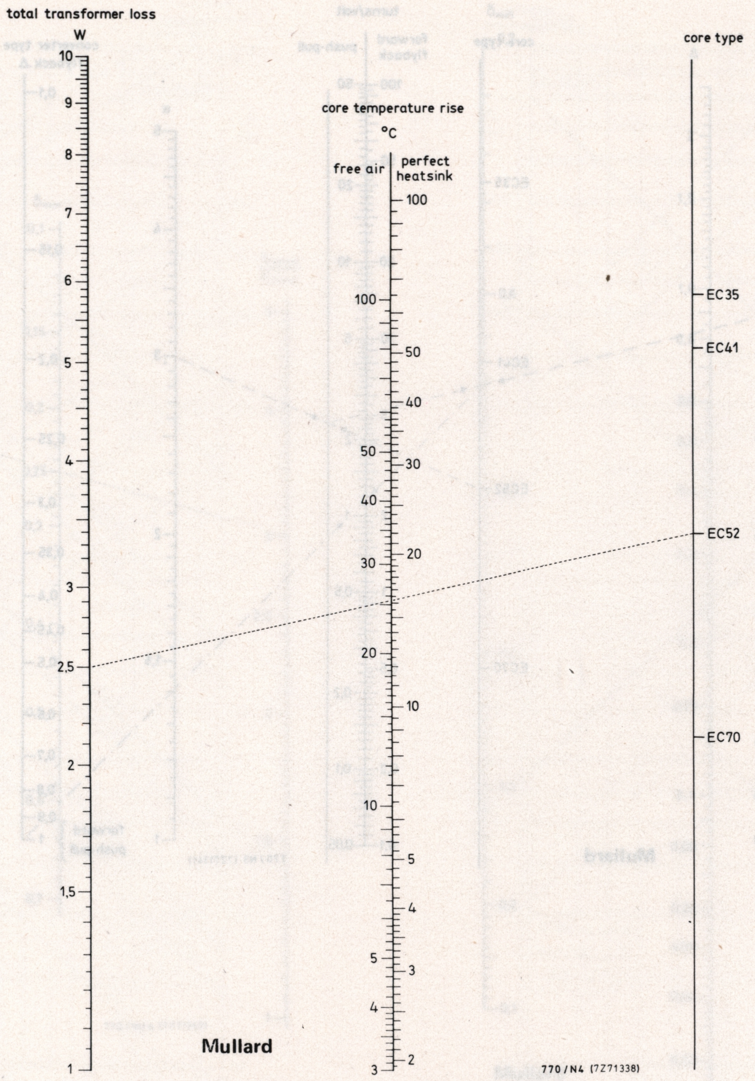
Nomogram 2 — Flyback converter: parameter Δ as a function of $I_{o(min)}/I_{o(max)}$

This nomogram is divided into two parts. In (a), the minimum duty cycle δ_{min} is obtained from δ_{max} and $V_{i(max)}/V_{i(min)}$; in (b), this value of δ_{min} is used with $I_{o(min)}/I_{o(max)}$ to derive Δ.


f kHz
core type
EC35
EC41
EC52
EC70
turns/volt
forward flyback
push-pull
α
converter type flyback Δ
forward push-pull
Mullard
770 / N3 (7Z71341)

Nomogram 3 — Winding turns per volt

Draw construction lines joining: core type and value of α; intersection with axis and frequency f. Then join the new intersection with the core type axis and either the converter type or, for a flyback converter, the value of Δ. Read the number of turns per volt given by the intersection of the latter line with the appropriate scale.

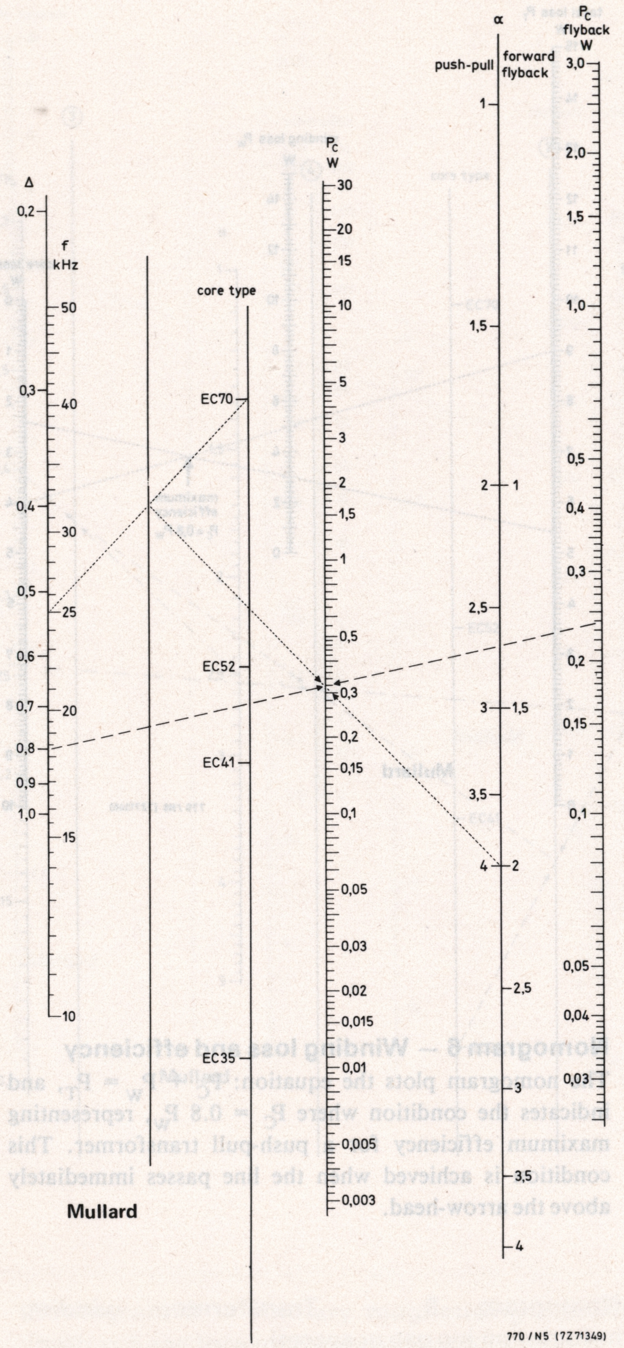

total transformer loss
W
core temperature rise
°C
free air
perfect heatsink
core type
EC35
EC41
EC52
EC70
Mullard
770/N4 (7Z71338)

Nomogram 4 — Total transformer loss, core type, and temperature rise

A straight line cuts the three variables.

The 'perfect heatsink' condition assumed is that the transformer core is clamped with one back face against a surface of constant temperature.

Range extension: increasing the total loss by a factor of 10^n increases the resulting temperature by 10^n.

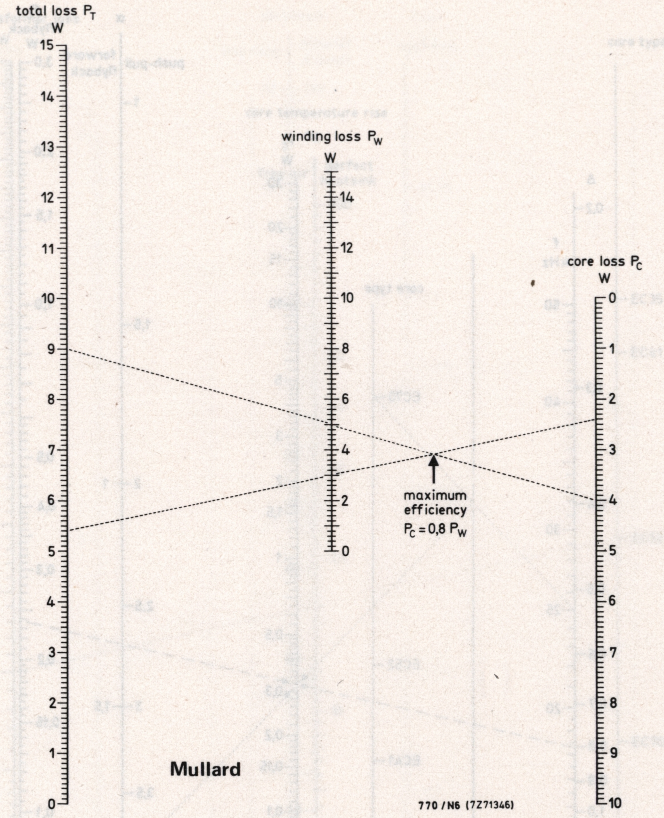

Δ
f
kHz
core type
EC70
EC52
EC41
EC35
P_C
W
α
push-pull
forward flyback
P_C flyback
W
Mullard
770/N5 (7Z71349)

Nomogram 5 — Core loss

Draw construction lines joining: frequency and core type; intersection with axis and the chosen value of α. The latter passes through the core loss P_C.

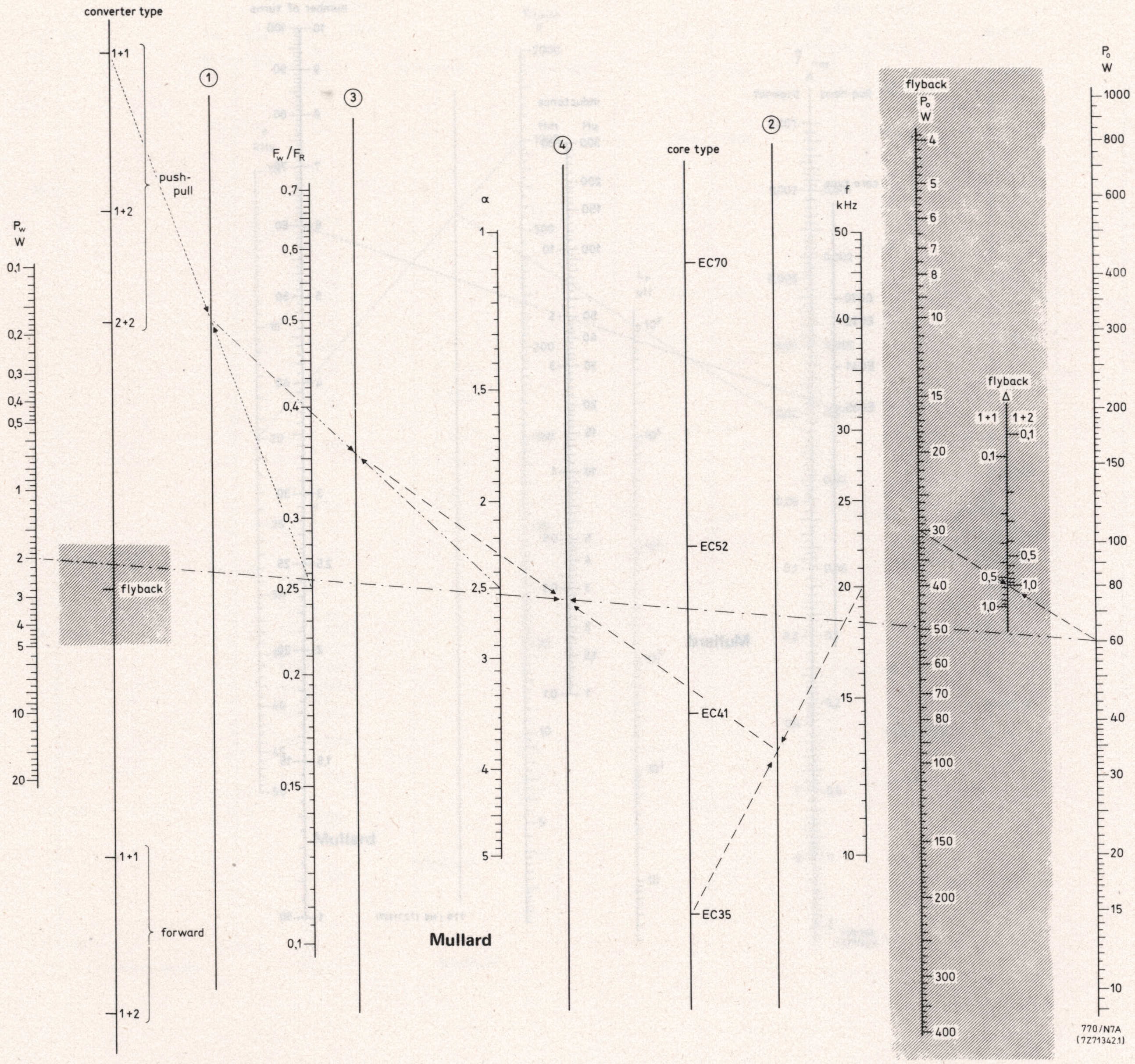
Note that two α scales are provided: one for push-pull converters, and the other for forward and flyback converters.

For a flyback converter the value of P_C must be modified using the parameter Δ. A straight line cuts the values of Δ, P_C, and the new core loss on the P_C (flyback) scale.


total loss P_T W
winding loss P_W W
core loss P_c W
maximum efficiency $P_c = 0{,}8\ P_W$
Mullard
770/N6 (7Z71346)

Nomogram 6 — Winding loss and efficiency

The nomogram plots the equation: $P_C + P_W = P_T$, and indicates the condition where $P_C = 0.8\ P_W$, representing maximum efficiency for a push-pull transformer. This condition is achieved when the line passes immediately above the arrow-head.


converter type
1+1
push-pull
1+2
2+2
flyback
1+1
forward
1+2
P_w W
①
F_w/F_R
③
α
④
core type
EC70
EC52
EC41
EC35
②
f kHz
flyback P_o W
flyback Δ
1+1
1+2
P_o W
Mullard
770/N7A (7Z71342.1)

Nomogram 7 — Transformer throughput power

Draw construction lines joining: converter type and the value of F_W/F_R; core type and frequency f. Then join the intersection with axis 1 and the value of α; the intersection with axis 2 and the intersection with axis 3. The intersection with axis 4 then gives the proportionality between winding loss P_W and throughput power P_o.

For a flyback converter transformer, the value of P_o must be modified. Draw a further line from P_o through the value of Δ appropriate to the winding configuration, to the P_o (flyback) scale.

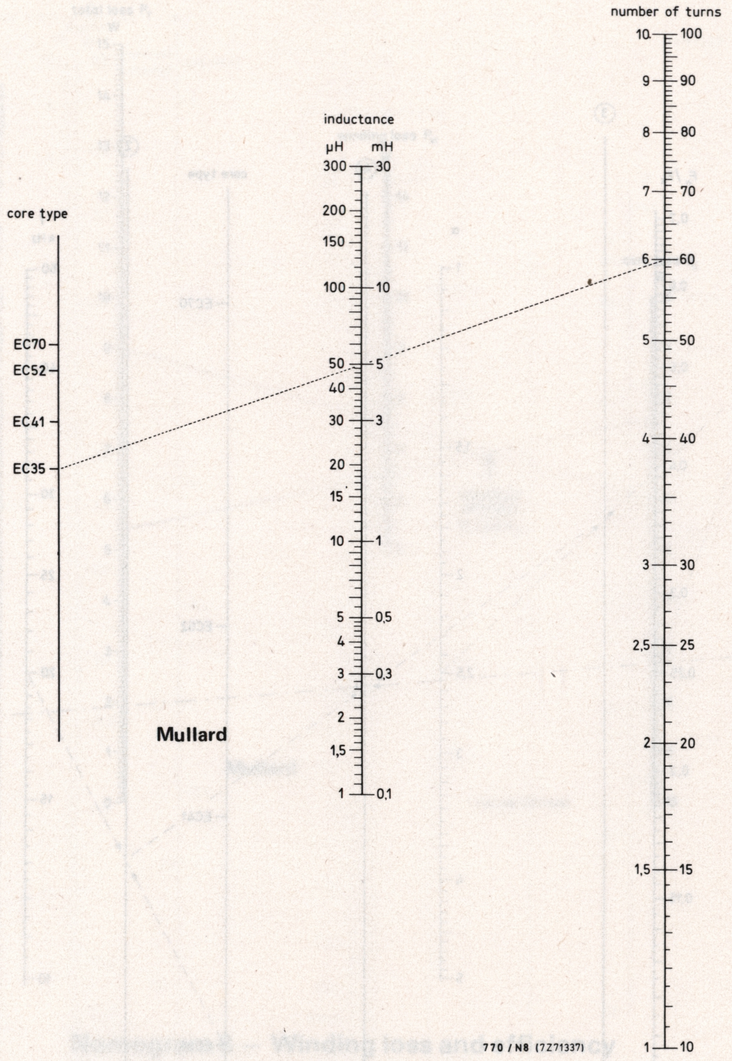
No range extension is possible within the permissible core operating conditions.

TABLE 3

Scaling Constants where 4mm Clearances are Omitted

$(P_o{}' = K_i P_o)$

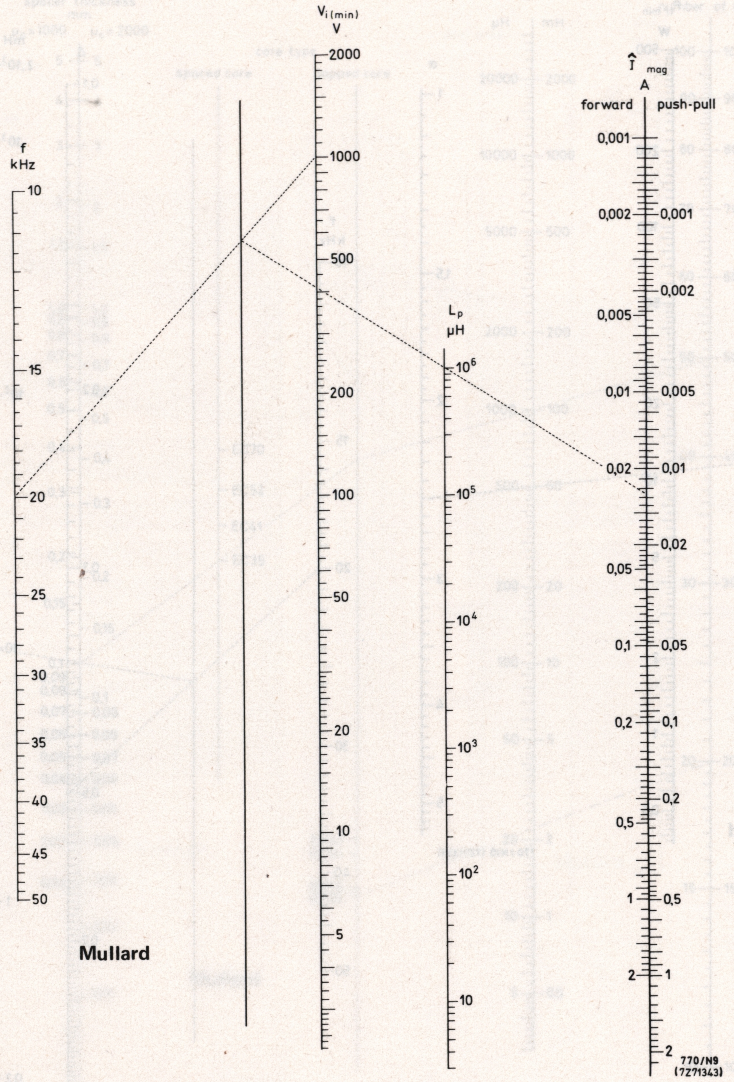
Core type	Ki
EC35	1.26
EC41	1.22
EC52	1.18
EC70	1.11



Nomogram 8 — Transformer winding inductance (no gap or spacer)

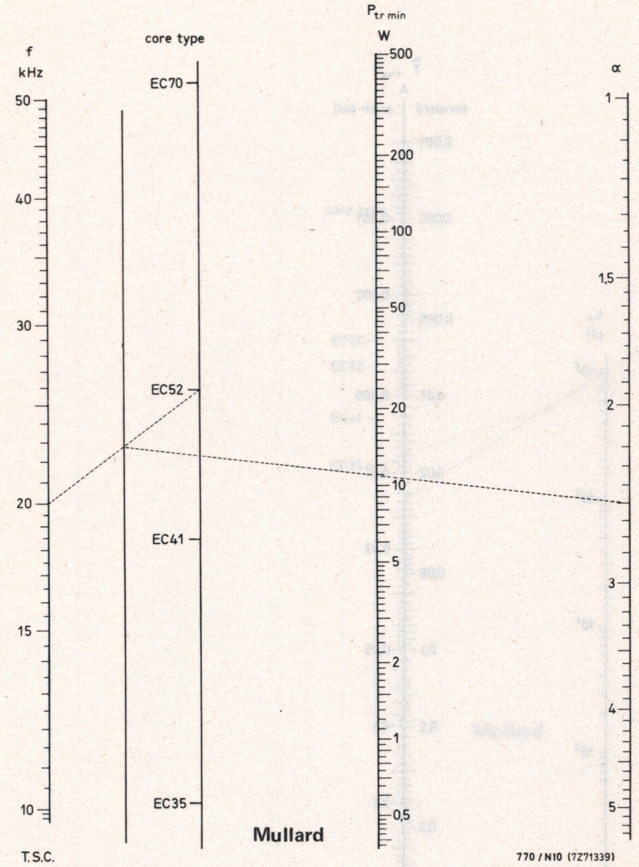
A straight line joins the core type, the inductance, and the number of turns.

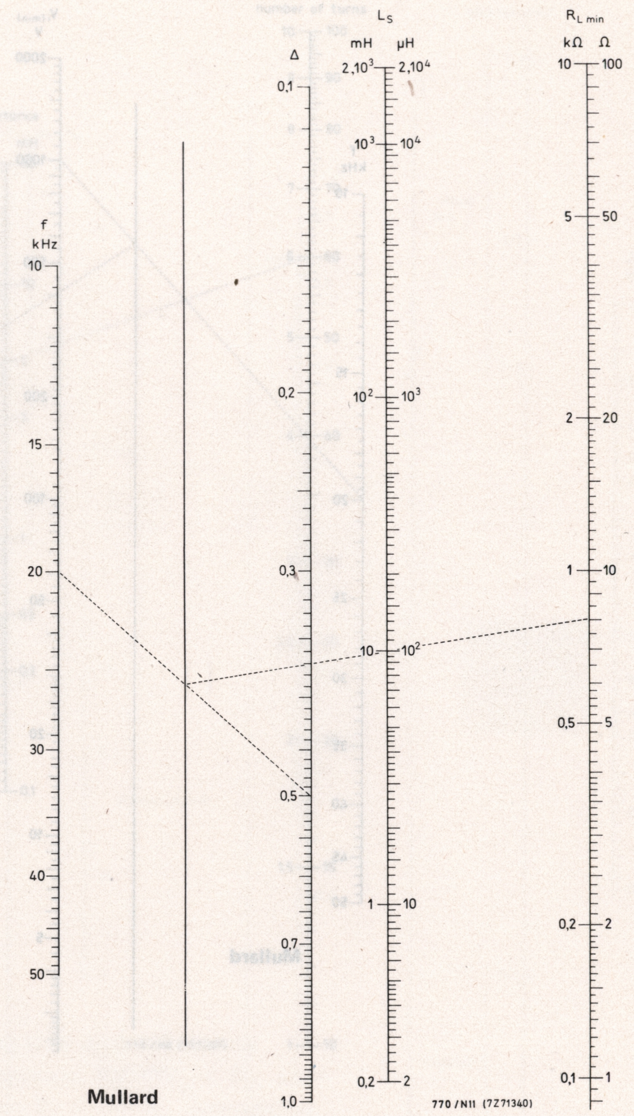
Range extension: changing the number of turns by a factor of 10^n changes the resulting inductance by 10^{2n}, as indicated.



Nomogram 9 — Primary magnetising current (push-pull and forward converter transformers)

Draw a construction line joining the frequency f and the minimum input voltage $V_{i(min)}$. A further line joins the intersection with the axis, the primary inductance L_p, and the peak magnetising current $\hat{I}_{mag}$.


f
kHz
50
40
30
20
15
10
core type
EC70
EC52
EC41
EC35
$P_{tr\,min}$
W
500
200
100
50
20
10
5
2
1
0,5
α
1
1,5
2
3
4
5
Mullard
T.S.C.
770/N10 (7Z71339)


f
kHz
10
15
20
30
40
50
Δ
0,1
0,2
0,3
0,5
0,7
1,0
L_S
mH
μH
$2,10^3$
$2,10^4$
10^3
10^4
10^2
10^3
10
10^2
1
10
0,2
2
$R_{L\,min}$
kΩ
Ω
10
100
5
50
2
20
1
10
0,5
5
0,2
2
0,1
1
Mullard
770/N11 (7Z71340)

Nomogram 10 — Minimum output power for good regulation (push-pull converters)

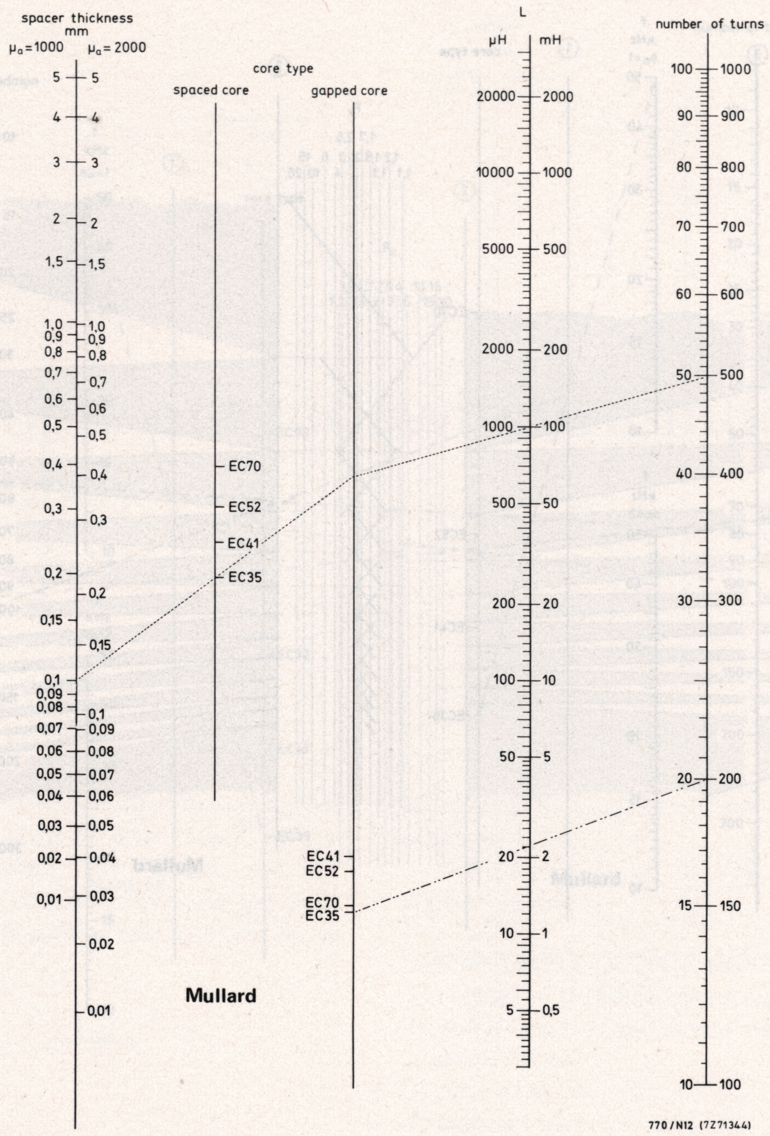
Draw a construction line joining frequency f and core type; a further line joins the intersection with the axis, the value of α, and the minimum throughput power $P_{tr(min)}$.

No range extension is possible within the permitted operating conditions.

Nomogram 11 — Secondary inductance for good regulation (flyback converters)

Draw a construction line joining the frequency f and the value of Δ. A further line from the intersection with the axis to the value of $R_{L(min)}$ passes through the value of L_s required for good regulation.

Range extension: changing the value of $R_{L(min)}$ by a factor of 10^n changes the resulting value of L_s by 10^{2n}, as indicated.


spacer thickness
mm
$\mu_a = 1000$
$\mu_a = 2000$
5
4
3
2
1,5
1,0
0,9
0,8
0,7
0,6
0,5
0,4
0,3
0,2
0,15
0,1
0,09
0,08
0,07
0,06
0,05
0,04
0,03
0,02
0,01
core type
spaced core
gapped core
EC70
EC52
EC41
EC35
EC41
EC52
EC70
EC35
L
μH
mH
20000
2000
10000
1000
5000
500
2000
200
1000
100
500
50
200
20
100
10
50
5
20
2
10
1
5
0,5
number of turns
100
1000
90
900
80
800
70
700
60
600
50
500
40
400
30
300
20
200
15
150
10
100
Mullard
770/N12 (7Z71344)

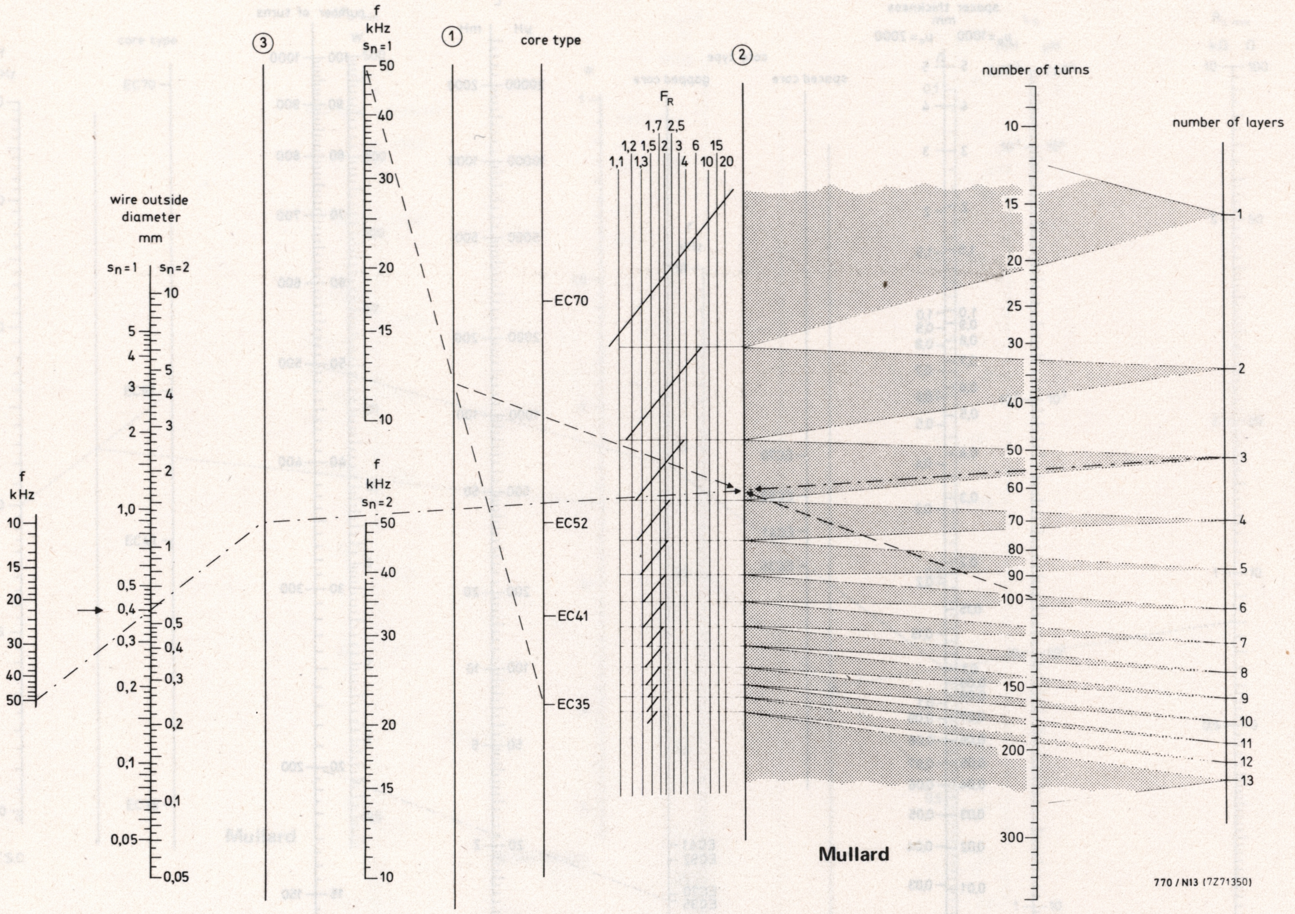
Nomogram 12 — Core spacer thickness or gap size (flyback converters)

Draw a construction line joining the number of turns, the inductance, and the gapped core axis. If the intersection with the latter axis is close to the marked point for the required core type, then this core with a standard gap may be used.

Otherwise, draw a further line from the intersection with the gapped core axis, cutting the required core type on the spaced core axis, and indicating on the spacer thickness scale the required spacer thickness.

Note that two permeability scales are provided for spacer thickness: the value $\mu_a = 1000$ is the minimum for the material.

At small spacer thicknesses (less than about 0.05mm), the practical problems associated with obtaining an accurate gap of such small dimension may make the calculated results difficult to reproduce.



Nomogram 13 – Copper wire diameter for minimum loss (single winding)

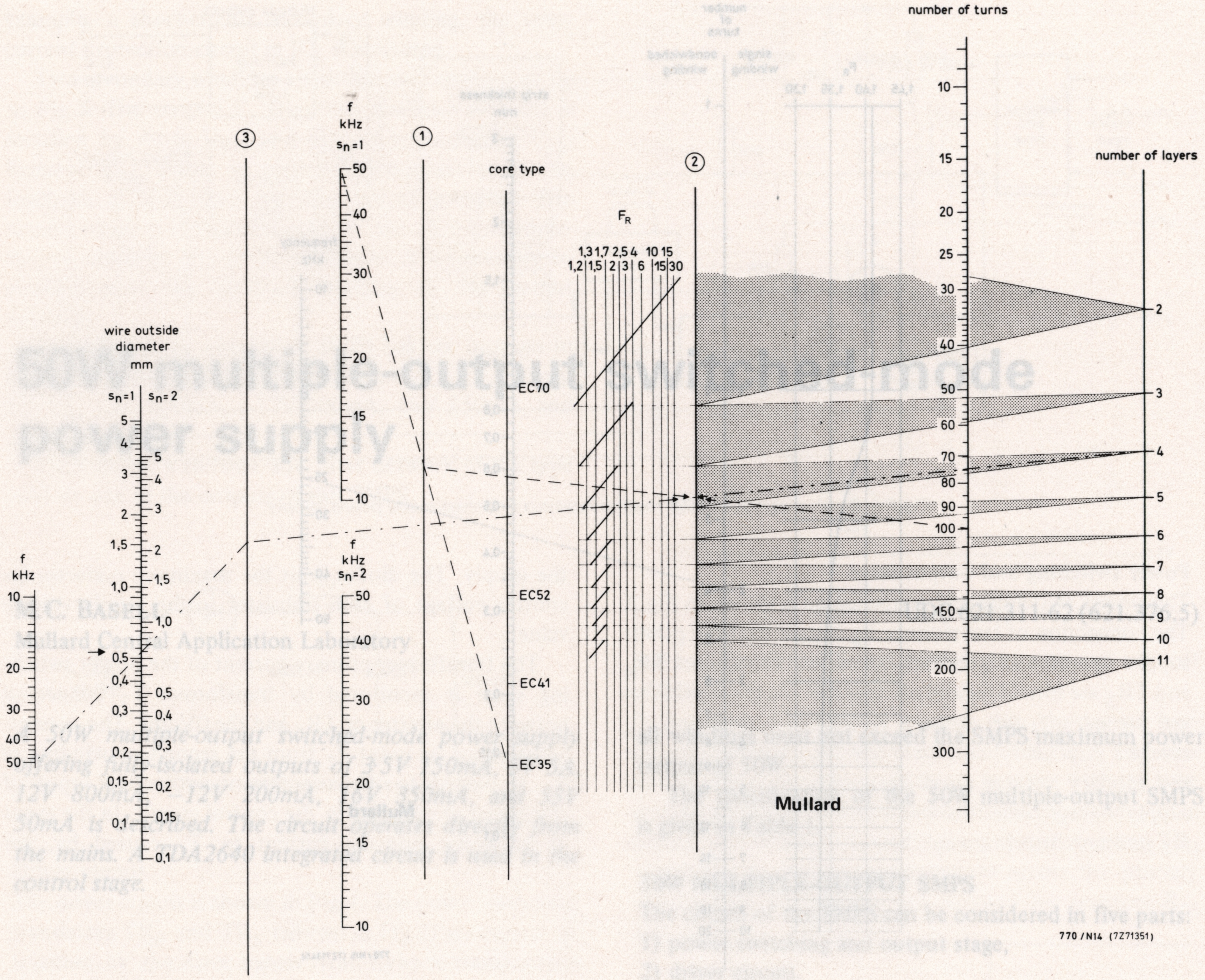
Draw construction lines joining: frequency f (for the appropriate number of strands) and core type; intersection with axis 1 and number of turns; intersection with axis 2 and the appropriate number of layers (follow shading). Finally join the intersection with axis 3 and the frequency f, to read the wire diameter from the appropriate scale for the number of strands s_n.

The value of F_R can be read from the graph adjacent to axis 2, by moving horizontally across from the intersection with this axis.

TABLE 5
Choice of Nomogram 13 or 14

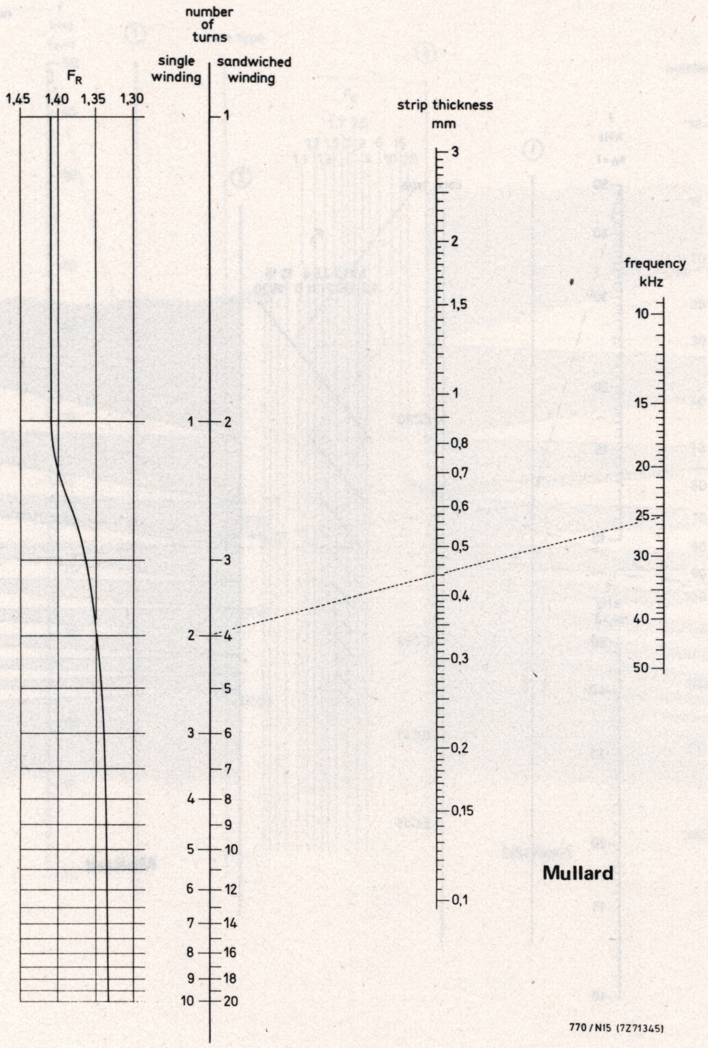
Winding type	Winding	Nomogram	n′	m′
Simple	Primary	13	n	m
Simple	Secondary	13	n	m
Split	Primary	13	n/2	m/2
Split	Secondary	14	n	m

n = total number of turns in the winding
m = total number of layers in the winding
n′ = number of turns to use in the nomogram
m′ = number of layers given by the nomogram



Nomograph 14 – Copper wire diameter for minimum loss (split or sandwiched windings)

Use as Nomogram 13.



Nomograph 15 — Required thickness of copper strip

Draw a straight line from the number of turns to the frequency: this line cuts the strip thickness scale at the optimum value for minimum loss.

Note that two 'number of turns' scales are provided. One is for a simple single winding, and the other is for a winding sandwiched between two others.

A graph of F_R as a function of the number of turns is also given.

50W multiple-output switched-mode power supply

M.C. Basell
Mullard Central Application Laboratory

UDC 621.311.62 (621.376.5)

A 50W multiple-output switched-mode power supply offering fully-isolated outputs of 3·5V 150mA, 5V 5A, 12V 800mA, −12V 200mA, 16V 350mA, and 55V 50mA is described. The circuit operates directly from the mains. A TDA2640 integrated circuit is used in the control stage.

INTRODUCTION

This 50W multiple-output switched-mode power supply uses a flyback converter and operates from the a.c. mains. A block diagram of the SMPS is shown in Fig.1. Each output of the SMPS is fully isolated. A TDA2640 integrated circuit is used in the control stage. Power for the control and driver stages is supplied from a feedback winding on the output transformer, and this feedback winding is also used for the regulation of the output voltages. A start-up circuit provides the initial supply for the control and driver stages when the mains is first applied to the SMPS. The current-sense circuit protects the power switching stage against excessive current in the primary of the output transformer caused by, for example, a short-circuit load on one of the secondary windings.

Any number of outputs can be provided as long as their associated windings will fit into the available winding space of the output transformer without the winding loss becoming too large. In addition, the total output of all windings must not exceed the SMPS maximum power output of 50W.

The specification of the 50W multiple-output SMPS is given in Table 1.

50W MULTIPLE-OUTPUT SMPS

The circuit of the SMPS can be considered in five parts:

1) power switching and output stage,
2) driver circuit,
3) control circuit,
4) transient start-up circuit,
5) input rectifier and filter circuit.

A brief description of each part is given in the following sections. The full circuit diagram comprising all five parts is given at the end of the descriptions (Fig.8).

TABLE 1

Specification of 50W Multiple-output SMPS

V_{in}	170 to 280V a.c.
P_{out} max	50W
P_{out} min	30W*
$f_{switching}$	25kHz
T_{amb}	0 to 60°C

*Minimum power output for good regulation

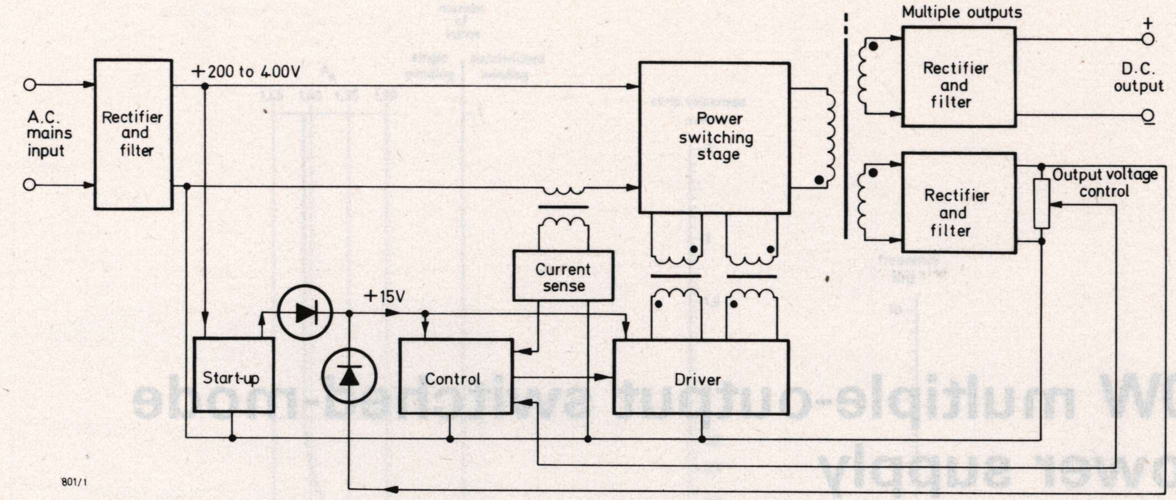


Fig.1 — Block diagram of 50W multiple-output switched-mode power supply

Power switching and output stage

The power switching and output stage is shown in Fig.2. The connection of transformer T_4 is such that the power supply operates as a flyback converter. Transistors TR_4 and TR_5 are turned on and off together. As the transistors are turned off, the primary voltage of transformer T_4 reverses in polarity, and the diodes D_{15} and D_{16} prevent the primary voltage from exceeding the value of the supply voltage. After the transistors turn off, the decreasing primary current passes through D_{15} and D_{16} back to the supply. The diodes limit the maximum collector-to-emitter voltage of each transistor to the supply voltage value.

The transformer primary current of the power switching stage is monitored by transformer T_3 connected between the emitter of TR_5 and the 0V line. The current-sense signal at point P_2 from the secondary winding of T_3 is passed to the overcurrent protection input of integrated circuit IC_1 in the control circuit. In the event of excessive primary current in T_4 caused by, for example, a short-circuit secondary winding, the overcurrent protection circuit in IC_1 is activated.

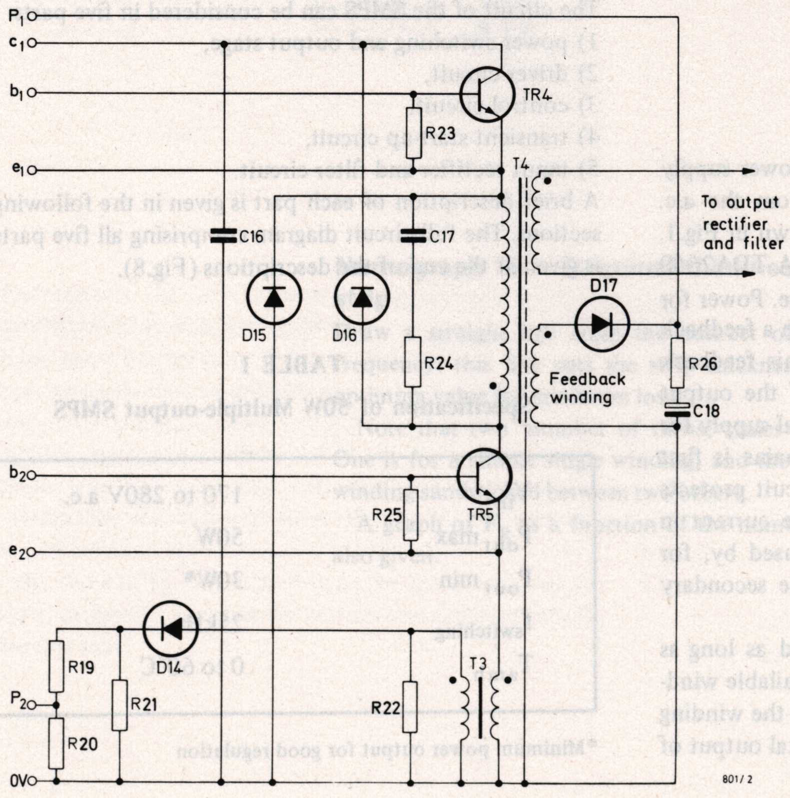


Fig.2 — Power switching circuit

Two secondary windings are shown on transformer T_4 in Fig.2. The output P_1 from the feedback winding is rectified by diode D_{17}, and filtered by capacitor C_{18}. Resistor R_{26} assists in reducing the phase-lag in the feedback system, thus ensuring overall stability. The second winding is representative of one of the multiple-output windings. For good regulation, it is necessary to closely-couple the feedback winding to the output windings.

The laboratory prototype SMPS has six outputs in addition to the feedback output, namely: 3·5V 150mA, 5V 5A, 12V 800mA, −12V 200mA, 16V 350mA, and 55V 50mA.

Fig.3 shows a rectifier and filter network of the type used on one of the outputs of this SMPS.

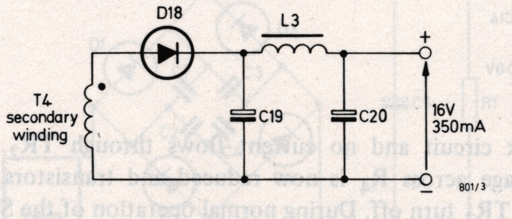


Fig.3 — Output rectifier and filter network

Driver circuit

The driver circuit is shown in Fig.4. The power switching transistors TR_4 and TR_5 (shown in Fig.2) are turned on together by the drive signals from flyback transformers T_1 and T_2. During the off-time of TR_4 and TR_5, transistor TR_3 is held in conduction by resistor R_{14}. Current builds up in the primaries of the driver transformers. and is limited by resistors R_{16} and R_{18}.

At the start of the on-time of TR_4 and TR_5, the control signal turns off TR_3 and prevents current flowing in the primary circuits of T_1 and T_2. As the flux in the transformer cores starts to fall, the primary voltages, and therefore the secondary voltages, reverse in polarity. The secondary voltages forward-bias the base-emitter junctions of TR_4 and TR_5, allowing current to flow in the secondary circuits, and providing the base current for TR_4 and TR_5. The stored energy in the transformers provides an almost-constant current source for the base of each power switching transistor.

At the end of the on-time of TR_4 and TR_5, the control signal turns on transistor TR_3. The collector voltage of TR_3 falls, and the primary and secondary voltages of T_1 and T_2 reverse in polarity, reverse-biasing the base-emitter junctions of TR_4 and TR_5. Thus both transistors are turned off.

Diodes D_{12} and D_{13} prevent currents circulating from one transformer to the other during turn-on and turn-off of TR_3.

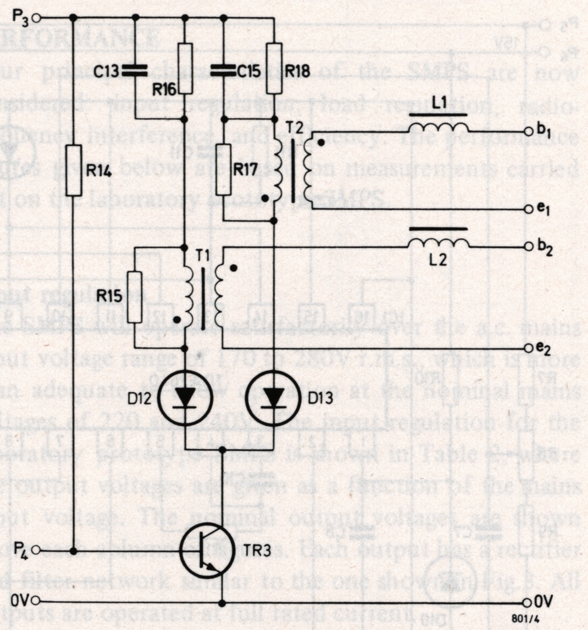


Fig.4 — Driver circuit

Inductors L_1 and L_2 control the removal of base stored charge during the turn-off of TR_4 and TR_5 (see Ref.1). These inductors which are in series with the base circuits of TR_4 and TR_5 prevent an instantaneous rise in secondary current in T_1 and T_2, and therefore when TR_3 switches off, the voltage transients across the primary windings of T_1 and T_2 are damped by resistors R_{15} and R_{17}.

Control circuit

The control circuit is shown in Fig.5. A TDA2640 integrated circuit IC_1 provides a fixed-frequency variable on-time signal to control the power switching transistors. A full description of the TDA2640 integrated circuit is given in Ref.2.

The nominal supply voltage for IC_1 is 12V, and this voltage is obtained from the 15V supply line by resistor R_{10} and voltage regulator diode D_{10}. The operating frequency of the SMPS is controlled by resistor R_{12} and capacitor C_{10}. The values of these components have been chosen so that the switching frequency is 25kHz, thus ensuring 'silent' operation of the SMPS. The open-collector output transistor of IC_1 sinks sufficient current to turn off the driver transistor TR_3 shown in Fig.4. The on-time of the power switching transistors corresponds to the on-time of the output transistor of IC_1.

The 16V rectified output signal from the feedback winding of transformer T_4 (see Fig.2) is used to regulate the output voltages of the SMPS by controlling the duty cycle. Resistors R_7, R_9, and potentiometer R_8 attenuate the feedback signal before it is applied to the control

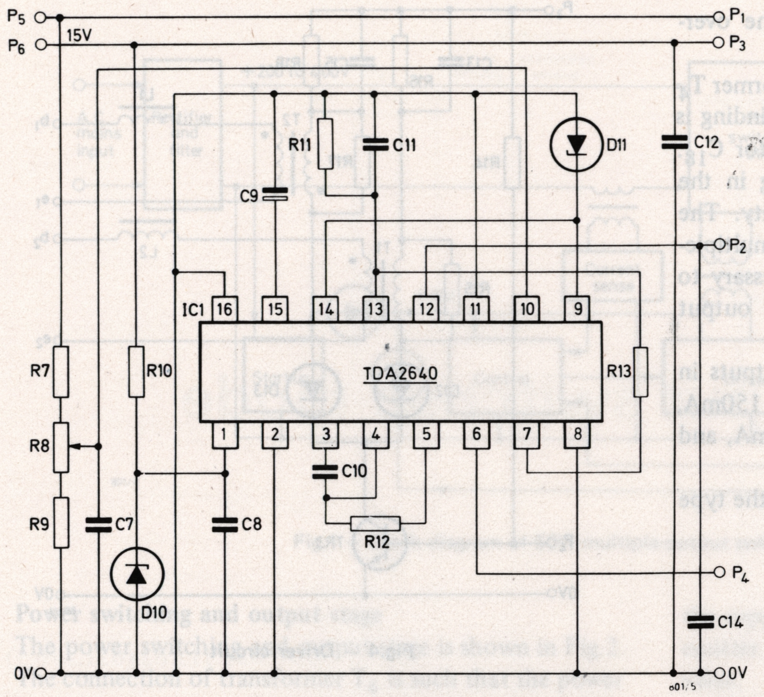


Fig.5 — Control circuit

input (pin 10) of IC_1. The attenuated signal is compared with the integrated circuit reference voltage of 6·2V. Overall adjustment of the output voltages of the SMPS is provided by potentiometer R_8.

The current-sense signal at point P_2 of the power switching stage is applied to pin 12 of IC_1. When excessive current in the primary of transformer T_4 is detected, IC_1 switches off the drive signal for approximately 1s before trying to restart the drive. When the overcurrent condition is removed, the circuit will restart.

Transient start-up circuit

The supply voltage for the control and driver stages is obtained from the feedback winding on transformer T_4. However, because this winding cannot provide an output until the control and driver stages are operating, an additional supply is required for these stages when the mains is first applied to the SMPS. The transient start-up circuit shown in Fig.6 provides this initial supply voltage.

Capacitor C_6 is charged through resistor R_2 when the mains is turned on. When the capacitor voltage reaches about 35V, the diac D_6 is triggered, turning on TR_1. Transistor TR_1 turns on transistor TR_2 and current flows from C_6 through R_4, TR_2, and D_9 to the control and driver stage supply circuit, thus starting the whole SMPS. The voltage regulator diode D_5 keeps the base of TR_2 at 15V, and therefore the supply to the control and driver stages is about 13·6V. Because only a small proportion of the current from C_6 passes through the collector of TR_1, as little as possible of the stored energy in C_6 is wasted.

When the feedback circuit voltage rises above 14·3V, the control and driver stages are supplied from the feedback circuit and no current flows through TR_2. The voltage across R_4 is now reduced and transistors TR_1 and TR_2 turn off. During normal operation of the SMPS, the voltage across C_6 is held almost constant at 35V by D_6.

Resistor R_2 has the rectified and smoothed mains voltage (less 35V) dropped across it when the SMPS is operating. To reduce the power dissipation, the value of R_2 must be as high as possible, but the charging time, and hence the time before the SMPS becomes fully operative, will increase as the value of R_2 increases. Therefore, a compromise value for R_2 has been chosen which gives a maximum power dissipation in the resistor of 2·6W, and a maximum delay before start-up of 7·5s.

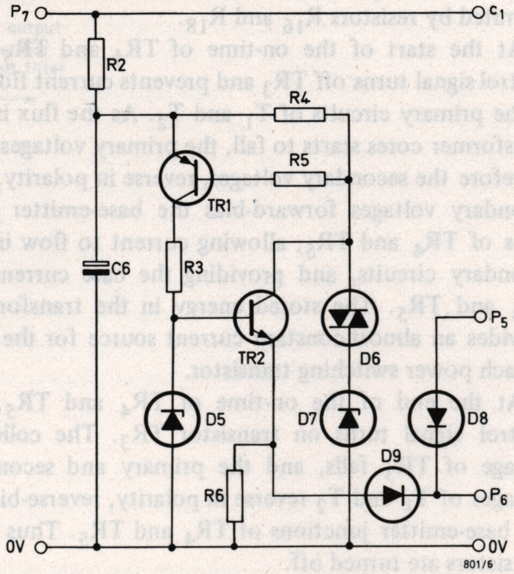


Fig.6 — Transient start-up circuit

The value of C_6 is chosen independently of R_2, and is determined on the basis of the maximum amount of energy required to start the SMPS.

Input rectifier and filter circuit

The input rectifier and filter circuit is shown in Fig.7. Diodes D_1, D_2, D_3, and D_4 are used as a bridge rectifier, and the d.c. output is filtered by smoothing capacitor C_5. Resistor R_1 assists in discharging C_5 after the SMPS is disconnected from the mains. An r.f. mains filter attenuates radio-frequency interference conducted to the mains supply.

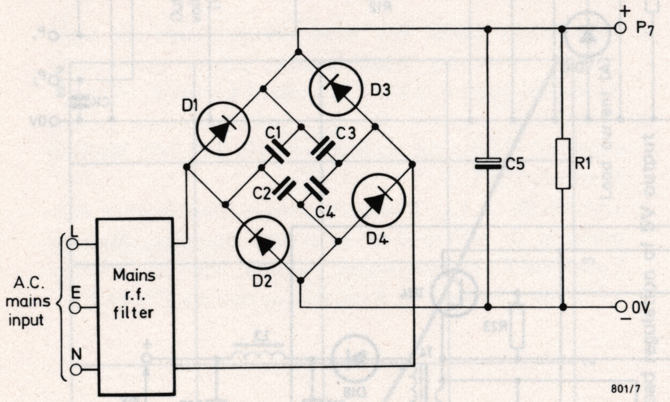


Fig.7 — Input rectifier and filter circuit

PERFORMANCE

Four principal characteristics of the SMPS are now considered: input regulation, load regulation, radio-frequency interference, and efficiency. The performance figures given below are based on measurements carried out on the laboratory prototype SMPS.

Input regulation

The SMPS will operate satisfactorily over the a.c. mains input voltage range of 170 to 280V r.m.s., which is more than adequate to allow operation at the nominal mains voltages of 220 and 240V. The input regulation for the laboratory prototype SMPS is shown in Table 2, where the output voltages are given as a function of the mains input voltage. The nominal output voltages are shown above each column of figures. Each output has a rectifier and filter network similar to the one shown in Fig.3. All outputs are operated at full rated current.

Load regulation

Load regulation curves for the six outputs of the laboratory prototype SMPS are given in Figs.9 to 14. For each curve, the output voltage is given as a function of the load current.

The load regulation of each output is largely dependent on the leakage inductance between each secondary winding and the feedback winding. In the laboratory prototype SMPS, the series resistance of the rectifier diode on the 5V output has a significant effect on the load regulation of this low-voltage output.

TABLE 2

Input Regulation of 50W Multiple-output SMPS

A.C. mains input voltage (V r.m.s.)	D.C. output voltages (V d.c.)					
	3·5V	5V	12V	−12V	16V	55V
180	3·37	5·16	11·89	−12·08	16·34	56·38
200	3·39	5·18	11·91	−12·09	16·38	56·51
220	3·40	5·20	11·93	−12·10	16·40	56·60
240	3·41	5·21	11·95	−12·11	16·43	56·69
260	3·42	5·22	11·96	−12·12	16·45	56·77
280	3·43	5·24	11·98	−12·13	16·47	56·85

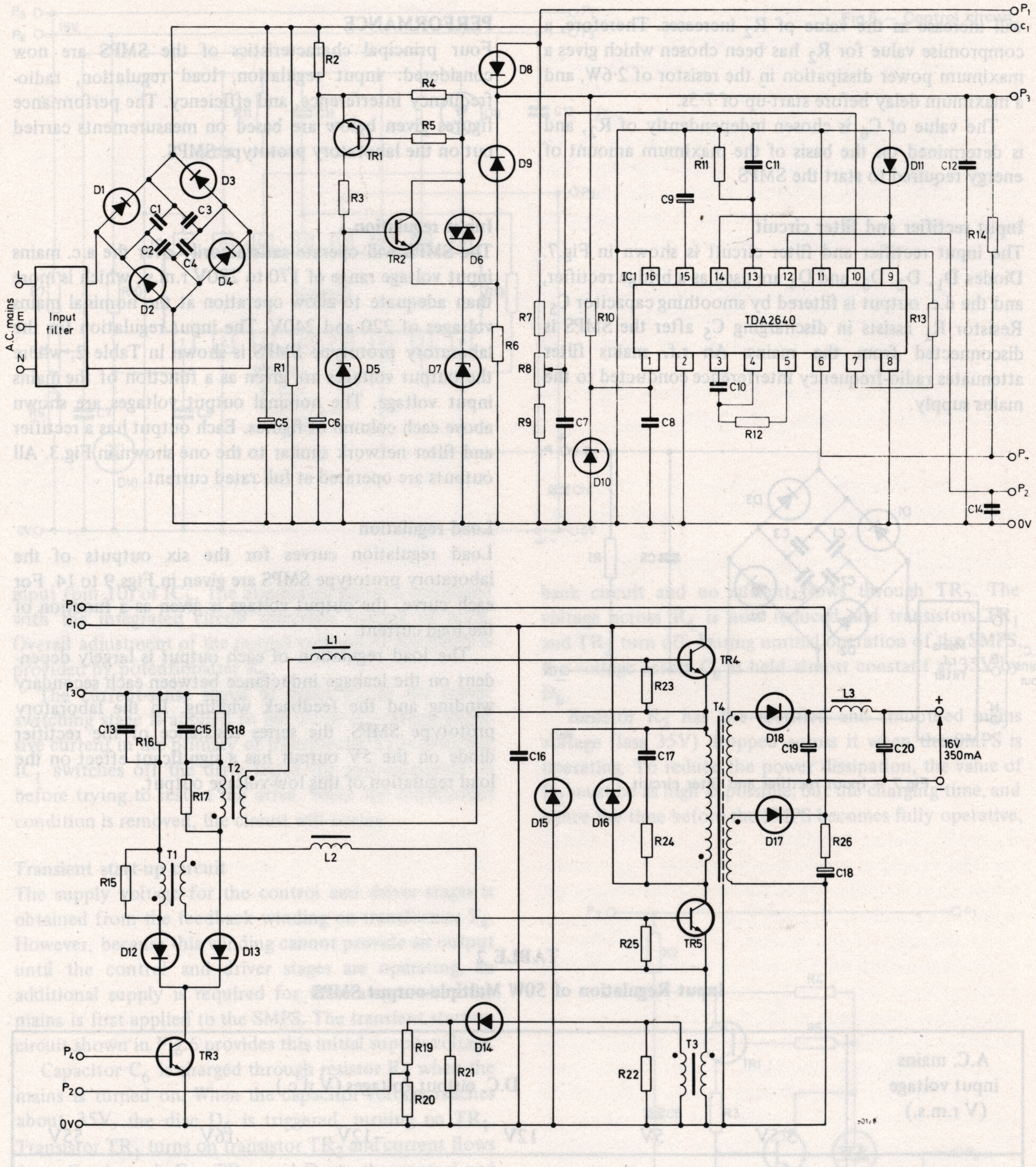

A.C. mains
L
E
N
Input filter
D1
D2
D3
D4
C1
C2
C3
C4
C5
C6
R1
R2
R3
R4
R5
R6
R7
R8
R9
R10
R11
R12
R13
R14
TR1
TR2
D5
D6
D7
D8
D9
D10
D11
C7
C8
C9
C10
C11
C12
C14
IC1
TDA2640
P1
c1
P5
P-
P2
0V
c1
C1
P3
C13
R16
C15
R18
T2
R17
L1
L2
T1
R15
D12
D13
P4
P2
0V
TR3
R19
R20
R21
D14
R22
T3
R23
TR4
T4
D15
D16
C16
C17
R24
R25
TR5
D17
D18
L3
C19
C20
16V 350mA
R26
C18

Fig.8 – Complete circuit of 50W multiple-output SMPS

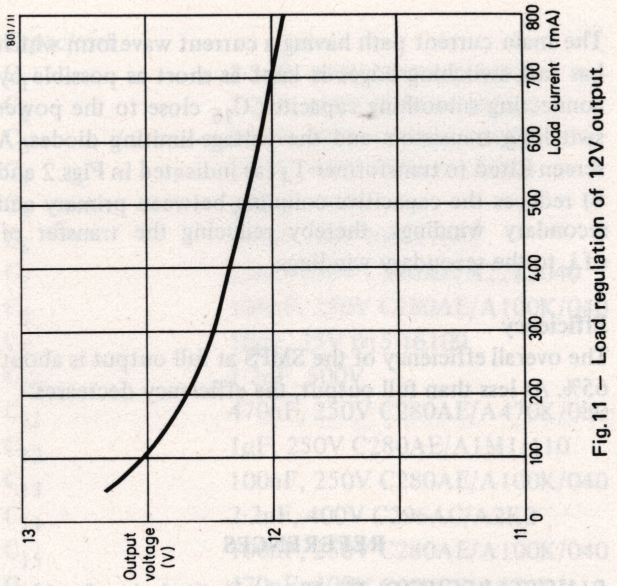


Fig.11 – Load regulation of 12V output

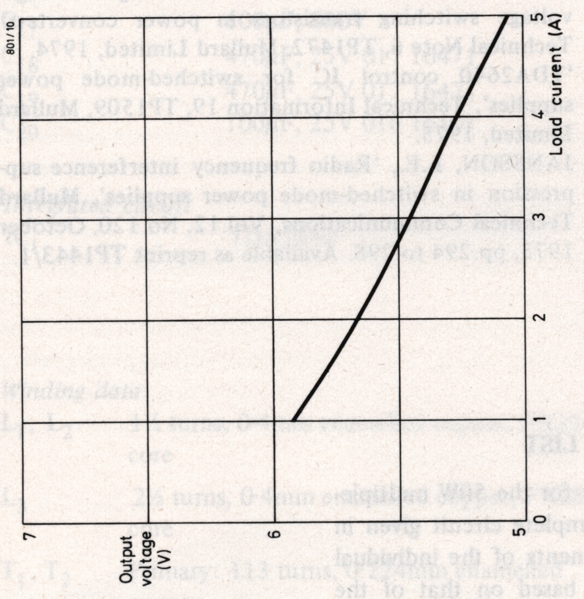


Fig.10 – Load regulation of 5V output

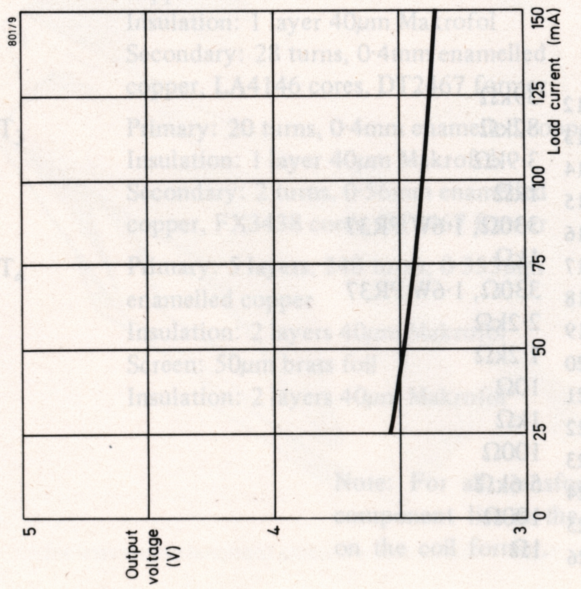


Fig.9 – Load regulation of 3·5V output

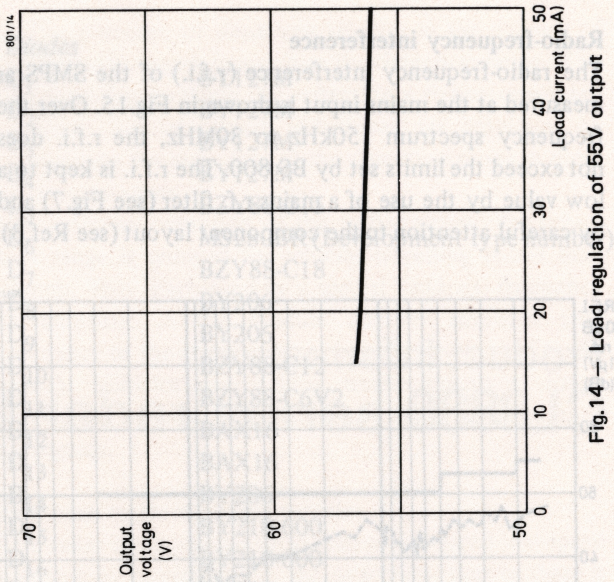


Fig.14 – Load regulation of 55V output

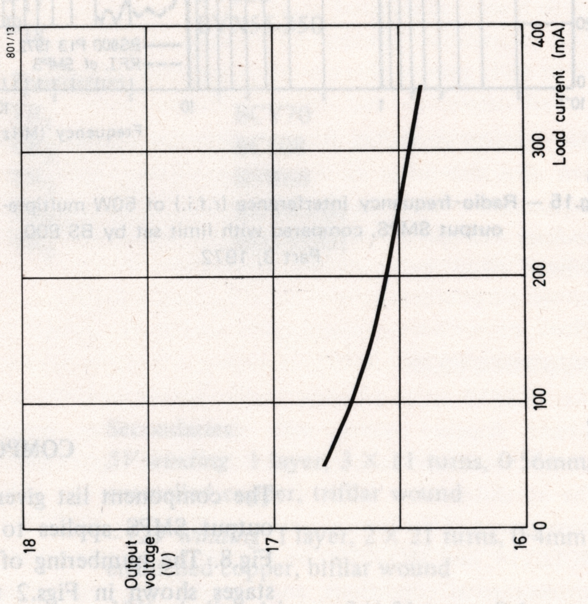


Fig.13 – Load regulation of 16V output

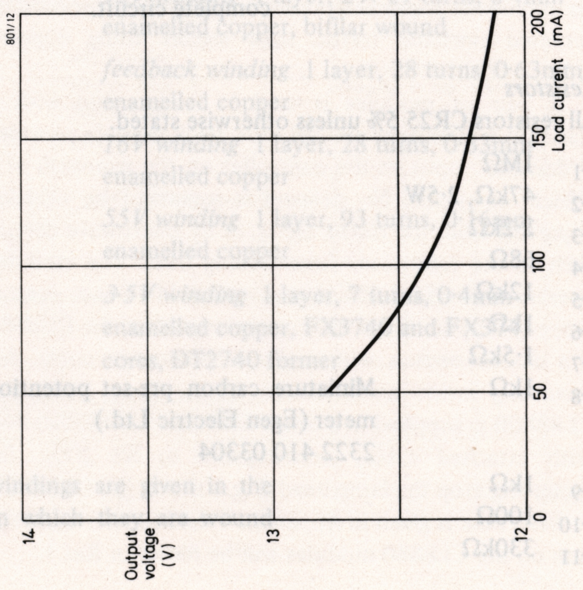


Fig.12 – Load regulation of −12V output

Radio-frequency interference

The radio-frequency interference (r.f.i.) of the SMPS as measured at the mains input is shown in Fig.15. Over the frequency spectrum 150kHz to 30MHz, the r.f.i. does not exceed the limits set by BS 800. The r.f.i. is kept to a low value by the use of a mains r.f. filter (see Fig.7) and by careful attention to the component layout (see Ref.3).

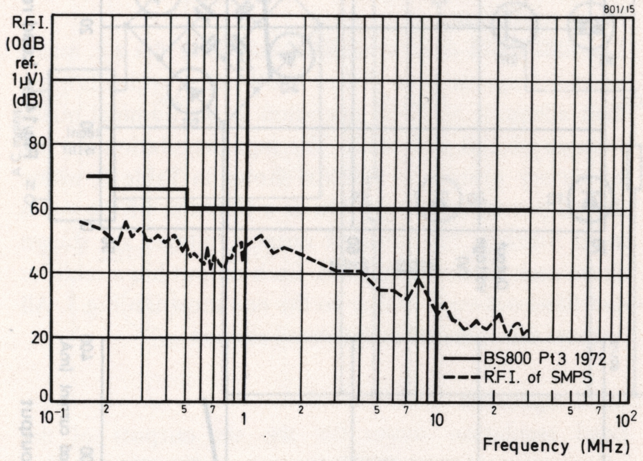


Fig.15 – Radio-frequency interference (r.f.i.) of 50W multiple-output SMPS, compared with limit set by BS 800, Part 3, 1972

The main current path having a current waveform which has fast switching edges is kept as short as possible by connecting smoothing capacitor C_{16} close to the power switching transistors and the voltage-limiting diodes. A screen fitted to transformer T_4 (as indicated in Figs.2 and 8) reduces the capacitive coupling between primary and secondary windings, thereby reducing the transfer of r.f.i. to the secondary windings.

Efficiency

The overall efficiency of the SMPS at full output is about 65%. At less than full output, the efficiency decreases.

REFERENCES

1. HETTERSCHEID, W., 'Base circuit design for high-voltage switching transistors in power converters', Technical Note 6, TP1472, Mullard Limited, 1974.
2. 'TDA2640 control IC for switched-mode power supplies', Technical Information 19, TP1509, Mullard Limited, 1975.
3. JANSSON, L.E., 'Radio frequency interference suppression in switched-mode power supplies', **Mullard Technical Communications**, Vol.12, No.120, October 1973, pp.294 to 298. Available as reprint TP1443/1.

COMPONENT LIST

The component list given below for the 50W multiple-output SMPS applies to the complete circuit given in Fig.8. The numbering of components of the individual stages shown in Figs.2 to 7 is based on that of the complete circuit.

Resistors

All resistors CR25 5% unless otherwise stated.

R_1	1MΩ	
R_2	47kΩ, 2·5W	
R_3	2·2kΩ	
R_4	68Ω	
R_5	12kΩ	
R_6	1kΩ	
R_7	1·5kΩ	
R_8	1kΩ	Miniature carbon pre-set potentiometer (Egen Electric Ltd.) 2322 410 03304
R_9	1kΩ	
R_{10}	100Ω	
R_{11}	330kΩ	

R_{12}	39kΩ
R_{13}	82kΩ
R_{14}	3·9kΩ
R_{15}	1kΩ
R_{16}	330Ω, 1·6W PR37
R_{17}	1kΩ
R_{18}	330Ω, 1·6W PR37
R_{19}	2·2kΩ
R_{20}	1·2kΩ
R_{21}	10Ω
R_{22}	1kΩ
R_{23}	100Ω
R_{24}	5·6kΩ
R_{25}	100Ω
R_{26}	1Ω

Capacitors

C_1	2·2nF, 400V C296AC/A2K2
C_2	2·2nF, 400V C296AC/A2K2
C_3	2·2nF, 400V C296AC/A2K2
C_4	2·2nF, 400V C296AC/A2K2
C_5	100μF, 400V
C_6	680μF, 40V 017 17681
C_7	22nF, 250V C280AE/A22K/040
C_8	100nF, 250V C280AE/A100K/040
C_9	10μF, 25V 015 16109
C_{10}	1·2nF, 500V
C_{11}	470nF, 250V C280AE/A470K/080
C_{12}	1μF, 250V C280AE/A1M1/110
C_{13}	100nF, 250V C280AE/A100K/040
C_{14}	2·2nF, 400V C296AC/A2K2
C_{15}	100nF, 250V C280AE/A100K/040
C_{16}	470nF, 400V C280CF/A470K/110
C_{17}	100pF, 800V
C_{18}	470μF, 25V 017 16471
C_{19}	470μF, 25V 017 16471
C_{20}	100μF, 25V 016 16101

Integrated circuit

IC_1	TDA2640

Winding data

L_1, L_2	1½ turns, 0·4mm enamelled copper, FX2249 core
L_3	2½ turns, 0·4mm enamelled copper, FX2837 core
T_1, T_2	Primary: 113 turns, 0·224mm enamelled copper Insulation: 1 layer 40μm Makrofol Secondary: 28 turns, 0·4mm enamelled copper, LA4146 cores, DT2467 former
T_3	Primary: 20 turns, 0·4mm enamelled copper Insulation: 1 layer 40μm Makrofol Secondary: 2 turns, 0·56mm enamelled copper, FX3438 cores, DT2467 former
T_4	Primary: 5 layers, 248 turns, 0·355mm enamelled copper Insulation: 2 layers 40μm Makrofol Screen: 50μm brass foil Insulation: 2 layers 40μm Makrofol Secondaries: *5V winding* 1 layer, 3 × 11 turns, 0·56mm enamelled copper, trifilar wound *–12V winding* 1 layer, 2 × 21 turns, 0·4mm enamelled copper, bifilar wound *12V winding* 1 layer, 2 × 21 turns, 0·4mm enamelled copper, bifilar wound *feedback winding* 1 layer, 28 turns, 0·63mm enamelled copper *16V winding* 1 layer, 28 turns, 0·63mm enamelled copper *55V winding* 1 layer, 93 turns, 0·16mm enamelled copper *3·5V winding* 1 layer, 7 turns, 0·4mm enamelled copper, FX3740 and FX3741 cores, DT2740 former

Diodes

D_1	BY127M
D_2	BY127M
D_3	BY127M
D_4	BY127M
D_5	BZY88-C15
D_6	MS280BR (Development type number)
D_7	BZY88-C18
D_8	BY206
D_9	BY206
D_{10}	BZY88-C12
D_{11}	BZY88-C6V2
D_{12}	BAX16
D_{13}	BAX16
D_{14}	BY206
D_{15}	BY210-600
D_{16}	BY210-600
D_{17}	BY206
D_{18}	BYX55-350

Transistors

TR_1	BCY70
TR_2	BC108
TR_3	BSW68
TR_4	BDY93/01
TR_5	BDY93/01

Note: For all transformers, windings are given in the component list in the order in which they are wound on the coil former.

Product Information:
CL8880 series radar traffic sensors

The modules described in this article use doppler radar with direction sense to detect approaching traffic. In conjunction with a traffic light control unit, a pair of modules will operate portable traffic signals according to demand. Other applications are possible.

INTRODUCTION

The CL8880 series of radar traffic sensor modules are X-band doppler radars, with direction sense, which provide vehicle actuation for portable traffic light signals. The modules have been designed to meet the U.K. Department of the Environment specifications MCE0111, MCE0114, and MCG0500 on technical performance and environmental conditions, and Home Office regulations on microwave transmitters contained in BR14 for vehicle-actuated portable traffic light signals using doppler radar.

A pair of modules is used in conjunction with a standard traffic light control unit. Each module provides a 'demand' input to the controller on the approach of a vehicle or cyclist. No demand is registered by vehicles departing from the traffic lights, or by microwave clutter such as the reflections from waving trees. The modules operate from 55-0-55V a.c. (that is, 110V a.c. centre-tapped) supplies. Two versions are available: the CL8880 is a chassis version with a plastic protective cover, for mounting inside an extension of the lamp head; the CL8880B is supplied in a heavy-duty cast aluminium box with attached cable, weatherproof to BS 2001, for mounting atop the lamp head. Fig.1 shows the CL8880B.

Brief data is given in Appendix 1, and further information on the CL8880 modules will be found in Ref.1.

The principal features of this equipment are as follows.

1) Good detection efficiency is achieved, and the detection range is in excess of the minimum specified by the Department of the Environment.
2) The equipment ignores the signal caused by departing vehicles.
3) The equipment detects lone vehicles approaching the traffic signals in the presence of large departing vehicles or a heavy stream of departing traffic. This is an important feature of the design philosophy of the radar process circuit.
4) The equipment does not give false outputs in the absence of moving targets.
5) Signals reflected from fluttering or waving trees at reasonable distances do not cause unwanted outputs, even if a module is aimed directly at the trees.
6) Built-in self-testing circuits ensure acceptable operation of the traffic signal system in the event of failure of the input supplies or a fault in the module itself.

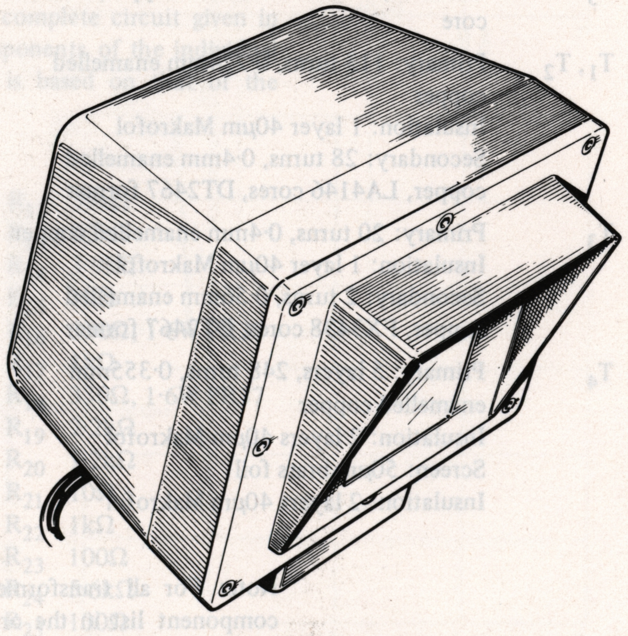


Fig.1 — Traffic sensor module (optional boxed version)

The CL8880 modules for vehicle-actuated traffic signals use a special type of doppler radar with direction sense. The units have several patented features, and result from many years' research and development in civilian and military microwave systems.

PERFORMANCE

The results of tests carried out on a test track with six radar traffic sensor modules are summarised in Fig.2, which shows the spread of measured detection range against vehicle speed. The minimum detection range (for small cars) specified by the Department of the Environment is shown for comparison. Four cars (Mini Traveller, Ford Cortina, Ford Escort, and Hillman Minx) were used in the tests on all six units. An additional test using an Aston Martin DB5 confirms the high-speed detection performance.

The tests were carried out under quiet road conditions. The detection range will be modified under the following operating conditions.

1) Vehicles such as lorries and buses, which have a large radar cross-section, will be detected at an increased range.
2) The presence of departing vehicles in the vicinity of the traffic signals can reduce the detection range of an approaching vehicle, by reflecting a strong unwanted signal. Special circuit techniques are employed to minimise the reduction of the detection range in these circumstances.

For optimum operation of the traffic signals, the detection range is reduced for vehicles travelling at less than 25mph. This is achieved by a special bandpass frequency characteristic in the module doppler amplifier (the frequency of the doppler signal being a function of target speed). Vehicles approaching traffic signals at lower speeds will not register a demand until they are nearer to the signals.

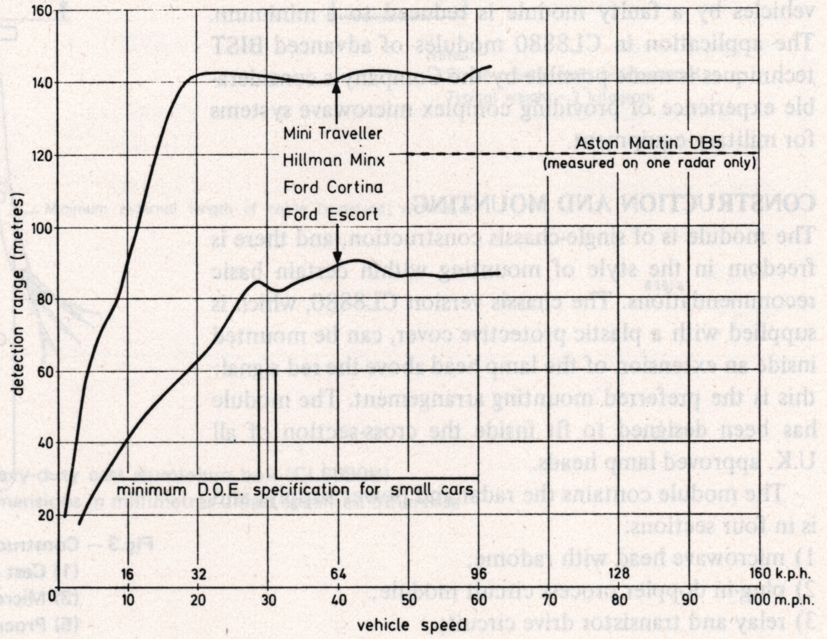
Response to slow-moving vehicles

The radar in its present form retains a reduced low-speed response to pedestrians in the near view; that is, there is no cut-off below 5mph. This is desirable because it allows a demand to be entered by 'inching' vehicles, cyclists, and vehicles approaching from a side turning close to the lights. These vehicles might not otherwise have a sufficient component of velocity in the forward direction (towards the traffic signal) to enter a demand.

OUTPUT

On detection of oncoming traffic, a relay in the module is de-energised and a pair of isolated contacts are opened to signify detection. The contacts remain open during the detection of a stream of traffic. The minimum open time is 1s (nominal).

Fig.2 — Detection curves obtained from road tests using six different radar modules



Should the supply voltage to the module be interrupted, then the module relay will be permanently de-energised and a continuous demand will be applied to the controller. The traffic signals will then automatically operate in a fixed-time manner.

Built-in self-testing

Built-in self-testing (BIST) circuitry in the module tests for certain fault conditions (internal and external), and in the event of a fault will switch the system to pseudo-fixed-time operation by supplying a continuous demand to the traffic signal controller. The BIST circuit will register a fault if:

1) an incorrect voltage is applied to the module,
2) the 0V line becomes disconnected,
3) there is an incorrect operation of the Gunn and mixer diodes in the microwave head,
4) the first-stage amplifiers fail,
5) the Gunn bias fails,
6) the input supply voltages fall below the specified lower limit.

Under fault conditions the traffic signal controlled by the module will operate on a fixed-time basis, the duration of the 'green' period being determined by the demand signal from the other radar module. Should both the modules switch to permanent demand (which can occur if the supply voltages fall below the specified margin), then the system will revert to operation on the fixed-time, minimum-green-extension principle. Thus the performance of the system is well defined even under fault conditions.

The BIST concept improves the serviceability of the system, because the majority of fault conditions would result in a simulated fixed demand. The possibility of a permanent 'red/red' due to non-detection of approaching vehicles by a faulty module is reduced to a minimum. The application in CL8880 modules of advanced BIST techniques is made possible by the Company's considerable experience of providing complex microwave systems for military equipment.

CONSTRUCTION AND MOUNTING

The module is of single-chassis construction, and there is freedom in the style of mounting within certain basic recommendations. The chassis version CL8880, which is supplied with a plastic protective cover, can be mounted inside an extension of the lamp head above the red signal: this is the preferred mounting arrangement. The module has been designed to fit inside the cross-section of all U.K. approved lamp heads.

The module contains the radar and power supply, and is in four sections:

1) microwave head with radome;
2) plug-in doppler process circuit module;
3) relay and transistor drive circuit;
4) power supply circuit.

Fig.3 shows the construction of the module, and Fig.4 the external appearance of the boxed version. Connection to the chassis version is made at the terminal block numbered (8) in Fig.3.

ALTERNATIVE VERSIONS

A traffic sensor module which operates from a d.c. supply is also available. This module is identical to the CL8880, except that the supply voltage is 24V ± 20% d.c. The type numbers are CL8881 (chassis version) and CL8881B (boxed version).

For use in the U.K. the centre frequency of traffic sensor modules in the CL8880 series is 10·587GHz. However, other operating frequencies may be available on request to comply with national requirements in other countries.

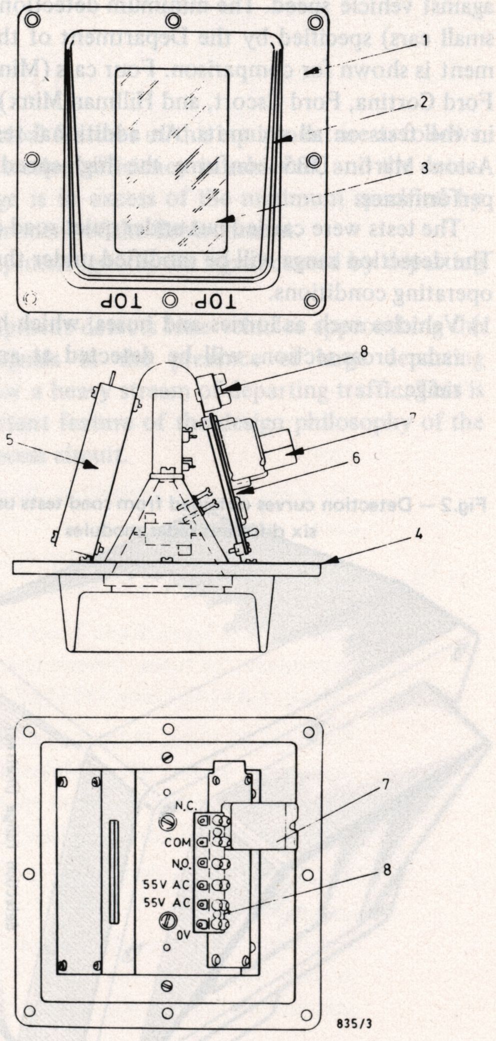


Fig.3 — Construction of CL8880 series modules:
(1) Cast aluminium front plate; (2) Cast aluminium cowl; (3) Microwave window (radome); (4) Rubber gasket seal; (5) Process circuit board; (6) Power supply circuit board; (7) Demand relay; (8) Terminal block

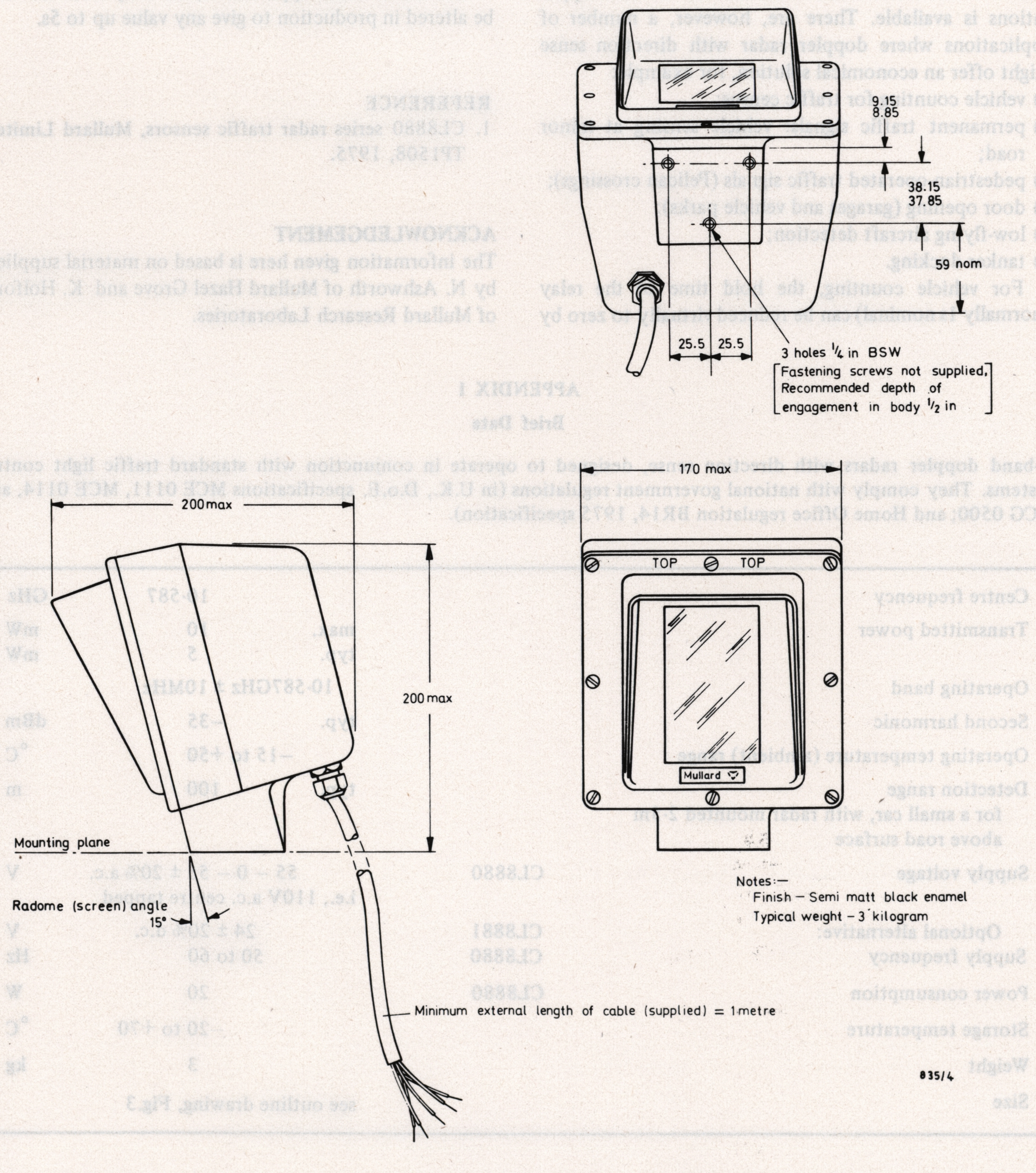

9.15
8.85
38.15
37.85
59 nom
25.5
25.5
3 holes ¼ in BSW
Fastening screws not supplied. Recommended depth of engagement in body ½ in
170 max
200max
200max
TOP ⊘ TOP
Mullard
Mounting plane
Radome (screen) angle 15°
Minimum external length of cable (supplied) = 1metre
Notes:—
Finish — Semi matt black enamel
Typical weight – 3 kilogram
835/4

Fig.4 — Heavy-duty cast aluminium box (CL8880B)
Dimensions in millimetres unless specified otherwise

OTHER APPLICATIONS

At present it is envisaged that CL8880 modules will be used only as sensors for vehicle-actuated traffic signals, and no detailed information on their use in other applications is available. There are, however, a number of applications where doppler radar with direction sense might offer an economical solution, for example:

1) vehicle counting for traffic census;
2) permanent traffic signals: vehicle sensing at minor road;
3) pedestrian-operated traffic signals (Pelican crossings);
4) door opening (garages and vehicle parks);
5) low-flying aircraft detection;
6) tanker docking.

For vehicle counting, the hold time of the relay (normally 1s nominal) can be reduced virtually to zero by changing a connection on the circuit board socket. The unit may then be pointed obliquely across the road to detect one or other of the traffic streams.

For use in other applications, the relay hold time can be altered in production to give any value up to 5s.

REFERENCE

1. CL8880 series radar traffic sensors, Mullard Limited, TP1508, 1975.

ACKNOWLEDGEMENT

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APPENDIX 1

Brief Data

X-band doppler radars with direction sense, designed to operate in conjunction with standard traffic light control systems. They comply with national government regulations (in U.K., D.o.E. specifications MCE 0111, MCE 0114, and MCG 0500; and Home Office regulation BR14, 1975 specification).

Centre frequency			10·587	GHz
Transmitted power		max.	10	mW
		typ.	5	mW
Operating band			10·587GHz ± 10MHz	
Second harmonic		typ.	−35	dBm
Operating temperature (ambient) range			−15 to +50	°C
Detection range for a small car, with radar mounted 2·5m above road surface		typ.	100	m
Supply voltage	CL8880		55 − 0 − 55 ± 20% a.c. i.e., 110V a.c. centre tapped	V
Optional alternative:	CL8881		24 ± 20% d.c.	V
Supply frequency	CL8880		50 to 60	Hz
Power consumption	CL8880		20	W
Storage temperature			−20 to +70	°C
Weight			3	kg
Size			see outline drawing, Fig.3	